N74-14890

OPTIMIZATION IN THE DESIGN OF A 12 GIGAHERTZ LOW COST GROUND RECEIVING SYSTEM FOR BROADCAST SATELLITES - VOLUME I - SYSTEM DESIGN, PERFORMANCE, AND COST ANALYSIS

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October 1972

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NASA CR-121184 Vol. I

(NASA-CR-121184-Vol-1) OPTIMIZATION IN THE DESIGN OF A 12 GIGAHERTZ LOW COST GROUND RECEIVING SYSTEM FOR BROADCAST SATELLITES. VOLUME 1: SYSTEM DESIGN, (Stanford Univ.) ~~174~~ 168 p HC CSCL 17B G3/07 N74-14890 Unclas 26821

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Volume I

System Design, Performance, and Cost Analysis

by

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Center for Radar Astronomy
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NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

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Final Report

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ABSTRACT

The technical and economical feasibility of using the 12 GHz band for broadcasting from satellites were examined. Among the assigned frequency bands for broadcast satellites, the 12 GHz band system offers the most channels. It also has the least interference on and from the terrestrial communication links.

The system design and analysis are carried out on the basis of a decision analysis model. Technical difficulties in achieving low-cost 12 GHz ground receivers are solved by making use of a die cast aluminum packaging, a hybrid integrated circuit mixer, a cavity stabilized Gunn oscillator and other state-of-the-art microwave technologies for the receiver front-end. A working model was designed and tested, which used frequency modulation. A final design for the 2.6 GHz system ground receiver is also presented.

The cost of the ground-terminal was analyzed and minimized for a given figure-of-merit (a ratio of receiving antenna gain to receiver system noise temperature). The results were used to analyze the performance and cost of the whole satellite system.

CHAPTER I

OBJECTIVE AND SUMMARY

1.1 PREFACE

This report consists of two volumes. Volume I discusses the system design, cost, and performance. Volume II presents details on antenna system design and RF interference analysis. These two volumes are organized such that they can be read and understood independently, offering convenience to readers of different interest and backgrounds.

1.2 BACKGROUND OF RESEARCH

Satellite communication systems have experienced rapid development over a period of barely one decade both in the International Telecommunications Satellite Consortium (INTELSAT) global system and the domestic system of the USSR, offering communications over great distances where terrestrial means could not compete on operational and economic grounds. Although both systems were mainly to provide common-carriers, namely, telephony or video channels to be integrated into the switching network, the technical development in the output power capability from satellites has enabled the satellite system to play a much broader role. One of the most promising areas is with broadcasting satellites, which would give enormous benefits to education, medical service, teleconferencing and other social development usages. The block diagram of the broadcasting satellite system is illustrated in Fig. 1-1.

In 1967, Stanford University[1] initiated the feasibility study of a home reception system of educational television programs from broadcasting satellites using tens of thousands of low-cost ground-terminals. After recognizing the promise of this study, Stanford University and

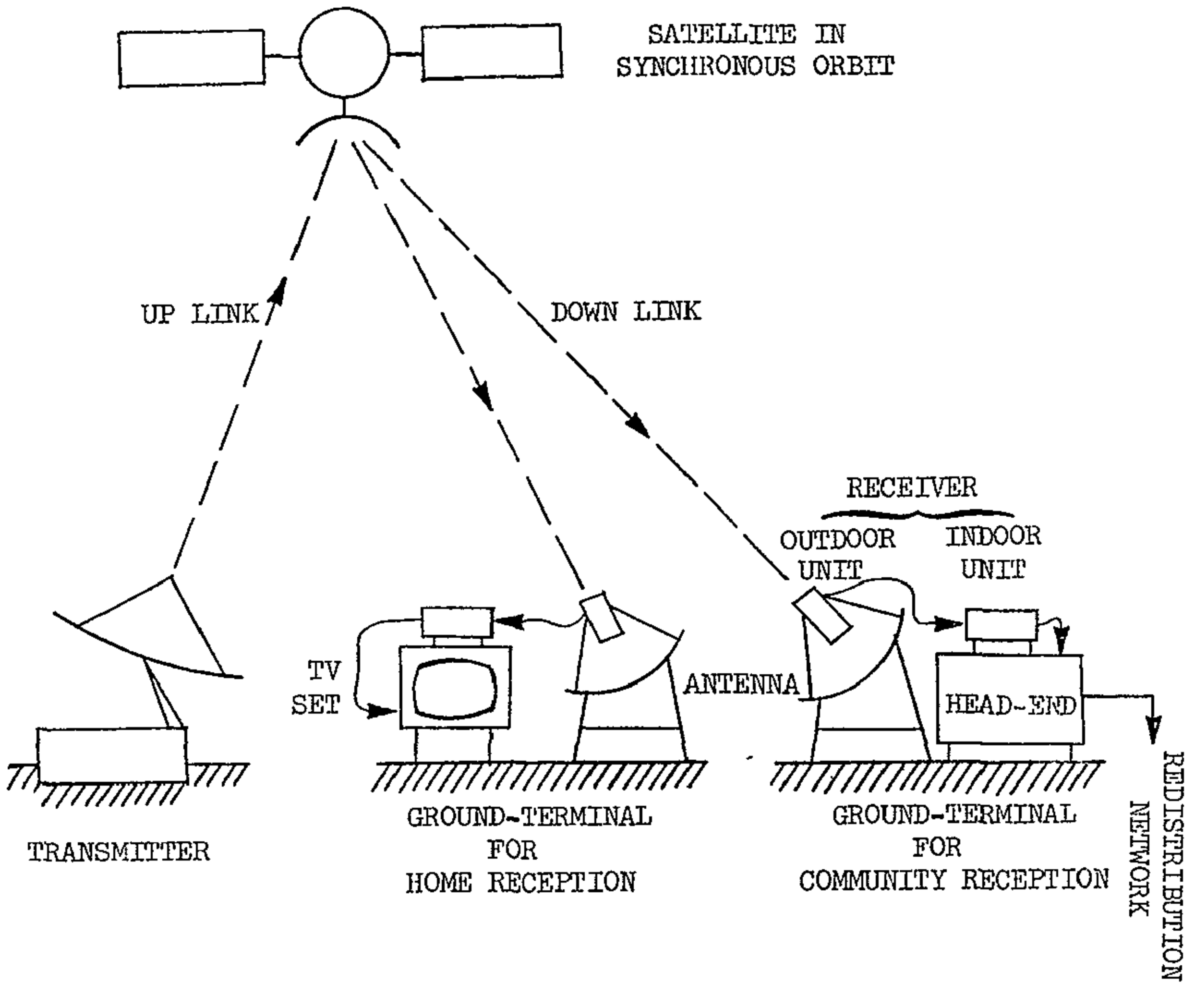


Figure 1-1 Block Diagram of Direct Reception from Broadcasting Satellite

National Aeronautics and Space Administration (NASA)[2] extended the study to design low-cost ground-terminals (antennas and receivers) suitable for mass-production and also calculated the communication system parameters for a 2.6 GHz system in 1971. In parallel, General Electric Company and NASA[3] designed and estimated the cost of mass-producible low-cost receivers (excluding antenna systems) for 2.6 and 12 GHz systems. These studies indicated great technical and economic potential of the 2.5 GHz direct reception system.

Although the practical potential of the 12 GHz system has not been clarified, some communication system parameters were calculated[4],[5] for frequency modulation (FM) and amplitude modulation (AM) systems. Utilization of the 12 GHz broadcasting satellite for the public broadcasting service and cable television (CATV) redistribution service (community reception) was studied and the communication system parameters were calculated by COMSAT[6] and Philco-Ford Corporation[7]. Ultra-high frequencies (UHF) were studied and prototype ground-terminals were designed for the community reception by Japan Broadcasting Corporation (NHK)[8].

In 1971, International Radio Consultative Committee (CCIR)[9] allocated three frequency bands, 620-790 MHz, 2.5-2.69 GHz and 11.7-12.2 GHz (for America, Asia and Oceania) or 11.7-12.5 GHz (for Europe and Africa), for the broadcasting satellite service and also regulated the maximum flux density on the earth for each frequency band. There are many problem areas involved with these three frequency bands: operational, economic and political. <u>The choice of a frequency band is one of the most important problems to be decided.</u>

The UHF band is least desirable for the direct reception system, especially for the home reception, because of the limited signal flux density allowed on the earth in order to avoid possible interference on the terrestrial UHF broadcasting services.

The 2.6 GHz band has been most thoroughly studied. Besides the studies mentioned above, J. G. Potter [10] analyzed the economic aspects of the teleconferencing networks using broadcasting satellites in this frequency band in order to minimize the annual system cost. Over and above these studies, an actual test of the 2.6 GHz direct reception from the ATS-F satellite to be launched in 1974 is being planned by the Federation of Rocky Mountain States, which will trigger the wider application of the broadcasting satellite in the near future.

The utility of the 12 GHz band for the direct reception, however, is less certain because of the difficulty in achieving low-cost mass-producible ground-terminals and the necessity in the system design to compensate for the signal attenuation caused by rain, although some link calculations and cost estimations were done[3],[4],[5] as mentioned before.

However, the current state of the art of microwave solid-state technology has enough promising future for the high-performance, low-cost and mass-producible 12 GHz receivers. Should the day come, the world would see practical exploitation of the 12 GHz broadcasting satellite system which would offer more channels and less interference problems than the other two alternative frequency bands.

1.3 OBJECTIVE

The purpose of this research is to solve technical problems on designing low-cost mass-producible 12 GHz ground-terminals, to design and test working models, to analyze and optimize the cost/performance of the 12 GHz system and to compare the 12 GHz band to the 2.6 GHz band. Topics are grouped into the following three categories:

Technical Aspects:

1. analyze frequency stabilization of solid-state microwave negative-resistance diode oscillators using a stabilizing cavity

2. use state-of-the-art microwave solid-state technology to achieve high-performance low-cost receivers in mass-producible way

3. integrate antenna subsystem in a way that the receiver outdoor unit packaging is die cast and the unit is installed at the focal point of the reflector

4. design and test working models

System Aspects:

5. apply decision analysis model to optimization in the design of receiver and communication systems

6. analyze and compare three modulation systems, AM, FM, and AM-FM
   the third modulation method, AM-FM, has not been noticed for the direct broadcasting system although the ground-terminal receiver can have a simpler configuration than the FM system

7. analyze the signal quality for the 2.6 GHz system and three communication system models of the 12 GHz system

Economic Aspects:

8. analyze and optimize the cost of ground-terminals and the whole system for the two frequency bands

9. analyze the system cost/performance sensitivity

10. compare the two alternative frequency bands

11. analyze the additional cost for multi-channel TV transmission

1.4 SUMMARY

Decision analysis provides a tool for systematically analyzing engineering problems. In Chapter II, models of the decision process are described which will be applied to all the decision processes throughout the paper. Communication system analysis for three alternative modulation techniques is discussed, showing that FM is the only practical method for the direct broadcasting. Time division multiplex is not considered here because of the higher cost of ground-terminals [10].

Chapter III describes the performance optimization and cost minimization of ground-terminal receiver design, using the decision model. The resulting optimum design and the test results of a working model are then discussed in Chapter IV. For clarity, these two chapters are organized such that all the lengthy analyses and design details are presented in appendixes rather than in the text. Appendix A describes a low cost technique of stabilizing the microwave negative resistance diode oscillator using invar cavity. The design of such a stabilized cavity is essential to the success of the design of a low-cost ground-terminal to be operated in stringent environments. Appendix B presents the decision analysis of the receiver subsystem whose results are used to make decisions on the selection of the main system. Appendix C summarizes all the details of the design of the working model's subsystems and components.

In Chapter V, system cost/performance is analyzed. First, the optimum combination of antenna diameter and receiver noise figure and the resulting minimum ground-terminal cost is given for a given ground-terminal figure of merit (a ratio of receiving antenna gain to receiver system noise temperature) and for a variety of production quantity per year. Second, the annual system cost is analyzed and minimized for a various number of ground-terminals. Cost/performance of low, medium, and high production quantities for the 12 GHz system is discussed, taking into account the effects of rain attenuation. Third, the actual signal-to-noise (SNR) ratio which subscribers obtain vs percentage of service outage is studied. Finally, it is shown that an appropriate 12 GHz system for the direct reception is about 60-80% more expensive than the 2.6 GHz system in terms of annual system cost, where the former has 15 hours a year of service outage during which the picture quality becomes below "passable" grade and 10 hours a year of outage during which the picture quality becomes below "inferior" grade if 12 hours a day of service time is assumed.

This report mainly discusses a single channel transmission system. However, practical applications to nation-wide education, medical service, teleconferencing, etc. would need more channels, especially for a redistribution system connected with terrestrial cable links. Modification of the receiver to multi-channel transmission is discussed in Appendix D. The additional receiver cost for up to 12 channel transmission is about 20%. Alternative modulation systems are also discussed for the multi-channel TV transmission.

In Chapter VI, the results of this study are reviewed and some future work is recommended.

CHAPTER II

COMMUNICATION SYSTEM ANALYSIS

2.1 MODEL OF DECISION PROCESS

Telecommunication system design is a decision process in which one must select the alternative with the best cost/performance measure among several possible choices at least from an engineering standpoint. In some cases the decision can be made on a deterministic basis but in many cases it must be done only on a stochastic basis. In the event, the decision for the optimization of the system design must be made in a systematic way. Hence, the methods of decision analysis provide a powerful tool for optimization in engineering problems.

The basic decision model by which our system design is to be built is shown in Fig. 2-1[11].

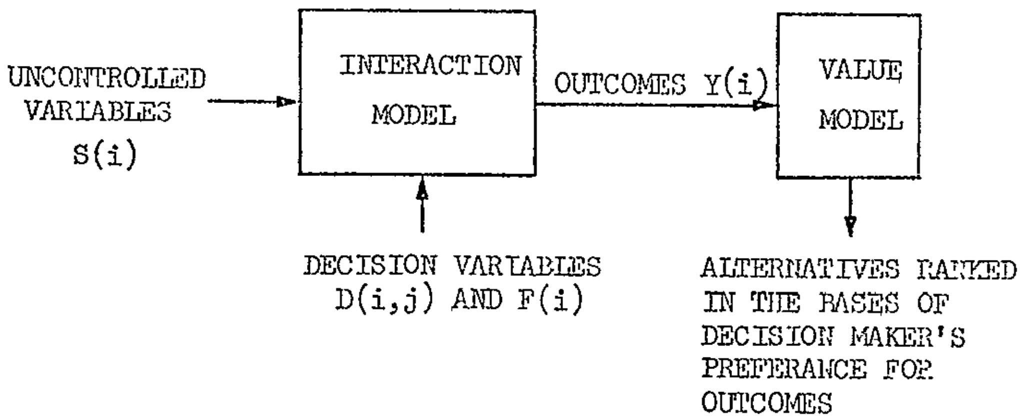


Figure 2-1 Model of Decision Process

In Fig. 2-1 there are two types of variables: (1) uncontrolled variables, which are variables not under the control of the decision maker; and (2) decision variables which are variables under the control of the decision maker, such as the system performance specifications.

The uncontrolled variables are expressed by a vector S(i), each component of which represents a constraint. Decision variables are grouped into two types: variables out of which we have to choose one discrete alternative------Matrix D(i,j), and variables which have a continuous value-----Vector F(i). An example of the former variable is the modulation method, AM or FM or AM-FM. If we regard i as the modulation method, we denote D(i,1), D(i,2) and D(i,3) as AM, FM and AM-FM method, respectively. That is to say, the row component shows the type of the variable and the column component shows its alternative. Examples of the latter variable F(i) are noise figure (NF) of the system, antenna aperture, and so on. We denote, for instance, F(1) and F(2) as NF and antenna aperture, respectively. The outcome is a vector Y(i), consisting of cost, performances, capability of future system modifications and so on.

Now we can extend this model to a system consisting of subsystems, which is shown in Fig. 2-2.

The type and the number of the subsystems are dependent on the choice of the decision variable D(i,j). That is to say, the subsystem would be different depending on whether we choose AM or FM or AM-FM, for example.

Each subsystem has its own uncontrolled variables $S_k(i)$ and decision variables $D_k(i,j)$ and $F_k(i)$. The uncontrolled variables of the whole system S(i) are of course uncontrolled variables of the subsystems and the decision variables of the whole system D(i,j) are also decision variables of the subsystems.

The output vectors of each subsystem are summed to give the outcome vector Y(i) of the system which is an input to the Value Model to rank the alternatives on the decision makers preferences.

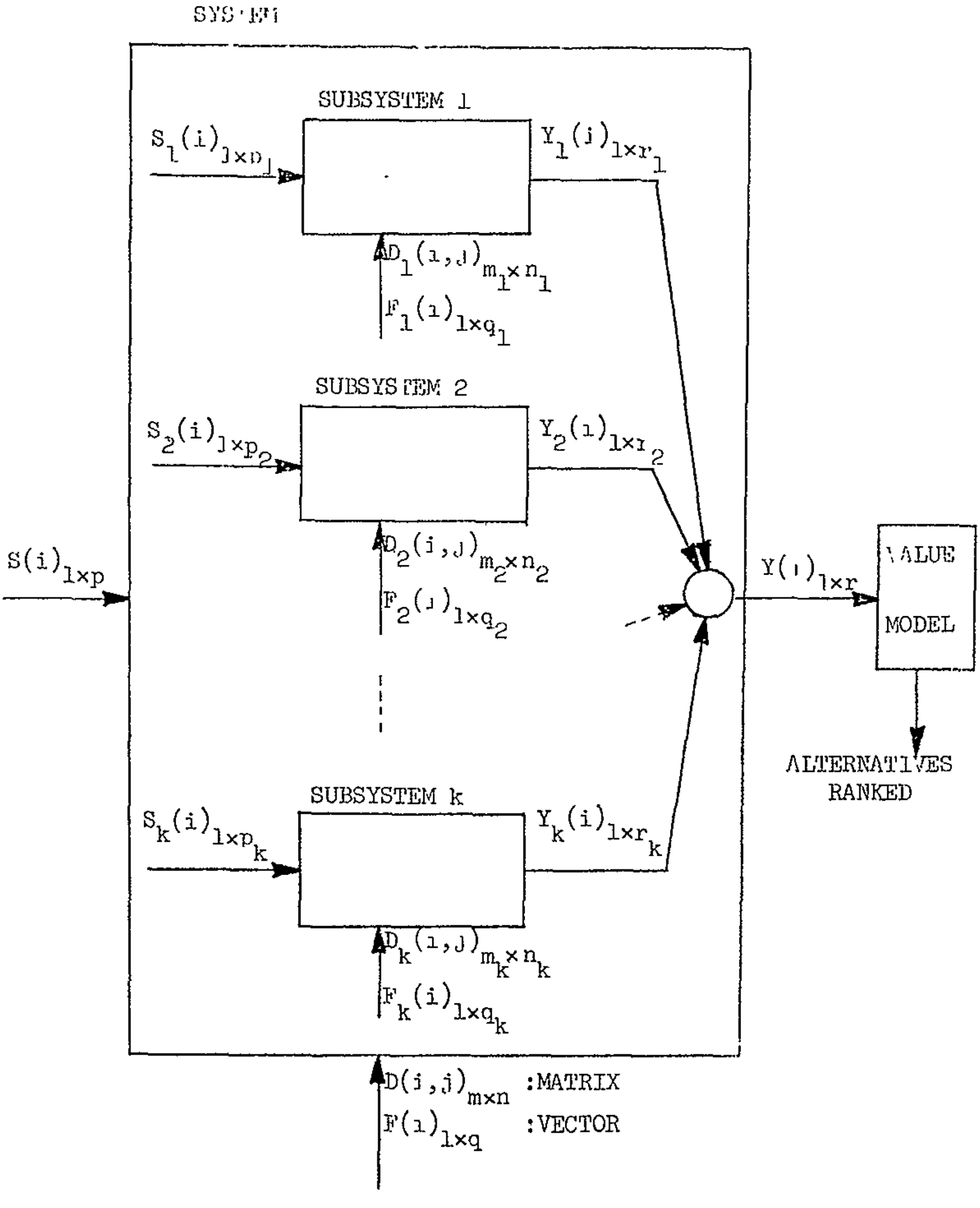


Figure 2-2 A Model of a Decision Process of a System Having Subsystems

This model is easily extended further into a more sophisticated system where subsystems have their own sub-subsystems.

This decision making procedure is very similar to the computer DO loop, where each DO loop can be regarded as a subsystem. Many calculations are indeed needed for this type of decision making. But when we can neglect the effect of a choice of a decision variable of the main system on a subsystem, the subsystem can be separated from the main system and the procedure becomes simple. For example, let us assume that the receiving system of broadcasting satellites is the main system and a local oscillator of a ground-terminal receiver is a subsystem. Some decision variables for the main system like a modulation method, polarization, ground-terminal SNR and so on do not affect the decision of the subsystem. Hence, the subsystem is separated from and independent of those decision variables of the main system.

In the discussion that follows, our system design will be carried out in a systematic way to achieve the best cost/performance by using the decision model.

2.2 MODULATION SYSTEM AND COMMUNICATION SYSTEM PARAMETERS

2.2.1 Picture Quality Considerations

In this chapter we analyze system parameters for three different modulation techniques - frequency modulation (FM), amplitude modulation (AM) and double modulation by AM and FM (AM-FM). These are the most likely alternative modulation techniques to be used for direct broadcast from satellites. Pulse techniques may also be used for

satellite television communication in the future but are excluded here because of the higher cost of the terminal receiver due to the complexity of its signal processing circuit[10].

The fundamental parameter of the communication system design is the signal-to-noise ratio (SNR). The desired SNR at the receiver output is not given and must be determined depending on the purpose of the system in order to give adequate picture and voice qualities at the terminal TV sets in some measurable sense. Generally, voice (audio) signals are far less subjective to noises than picture (video) signals, and so we limit the discussion to video signals hereafter. Here, SNR is defined as so called weighted SNR, which is the ratio of the peak to peak picture signal power to the rms noise power in the IF band multiplied by a noise weighting factor (NWF).

The relation between SNR and subjective (eye) picture quality was studied by Television Allocations Study Organization (TASO), 1960[12] and 1961, National Television System Committee (NTSC), 1958 and others[7]. The relation between the TASO grade and SNR is depicted in Fig. 5-11 in Chapter V.

Our purpose is to provide adequate signal quality both to the home receiving system (individual houses, schools, hospitals and so on) which has a small antenna and/or a moderate NF receiver and to the community receiving system (distributed for public broadcasting and cable TV) which has a larger antenna and/or a lower noise receiver. The latter needs a higher SNR at the receiver output than the former, because the latter distributes the received signals to a number of terminals. In this chapter we assume that signal quality must be guaranteed for more than 99% of

the service time[13]. A detailed sensitivity study will be done in Chapter V.

Now decide worst-case SNR as follows:

SNR for the home reception = 45 dB which corresponds to TASO/FCC grade 2

SNR for the community reception = 54 dB which corresponds to TASO/FCC grade 1.0.

The restriction set forth by CCIR (Geneva, 1971) [9] gives an upper bound on the flux density on the earth not to exceed -137 dBW/m^2 in any 4 KHz band for angles of arrival between 25 and 90 degrees above the horizontal plane for the 2.5 to 2.69 GHz band. This corresponds to the maximum effective isotropic radiated power (EIRP) from the satellite of 63 dBW for 27.5 MHz bandwidth. There is no limitation in EIRP for 11.7 to 12.2 GHz band. A 45 dB signal-to-noise ratio (SNR) for direct reception must be obtained by a low-cost ground station which consists of a relatively small and handy parabola and a low-cost receiver. This chapter analyzes the system parameters assuming the optimum model ground terminals shown in Table 2.1. The precise cost-performance sensitivity analysis with regard to specific hardware realization will be done in Chapter V.

Now, the decision model used to decide the best modulation system is summarized as follows:

Uncontrolled variables:

1. SNR $\geq$ 45 dB for home reception
   $\geq$ 54 dB for community reception
2. EIRP $\leq$ 63 dBW/27.5 MHz for the 2.5 to 2.69 GHz band
3. guaranteed service time $\geq$ 99%

Decision variables:

modulation system; AM, FM, AM-FM

Table 2-1

Model Ground Terminals

	2.6 GHz band	12.0 GHz band
Direct Reception ($SNR = 45$ dB)	$G_r/T_s = 0.4$ dB antenna diameter: 7 ft. $G_r = 33$ dB $B_{1\ dB} = 2.1^\circ$ preamplifier: none $NF_r = 8$ dB	$G_r/T_s = 11.6$ dB antenna diameter: 7 ft. $G_r = 46$ dB $B_{1\ dB} = .46^\circ$ preamplifier: none $NF_r = 10$ dB or antenna diameter: 5 ft. $G_r = 43$ dB $B_{1\ dB} = .65^\circ$ preamplifier: TDA $NF_r = 10$ dB
Community Reception ($SNR = 54$ dB)	$G_r/T_s = 9.4$ dB antenna diameter: 14 ft. $G_r = 39$ dB $B_{1\ dB} = 1.05^\circ$ preamplifier: transistor $NF_r = 5$ dB or antenna diameter: 11 ft. $G_r = 37$ dB $B_{1\ dB} = 1.3^\circ$ preamplifier: uncooled parametric $NF_r = 3$ dB	$G_r/T_s = 20.6$ dB antenna diameter: 10 ft. $G_r = 49$ dB $B_{1\ dB} = .31^\circ$ preamplifier: uncooled parametric $NF_r = 4$ dB

G_r/T_s = ground-terminal figure of merit = receiver antenna gain G_r divided by system noise temperature T_s

$B_{1\ dB}$ = 1 dB beam width

NF_r = receiver noise figure

55% of antenna efficiency is assumed

Outcomes:
1. bandwidth per channel
2. required EIRP
3. cost

2.2.2 Interaction Model - System Analysis
[A] FM System

A composite signal of a base-band video and a FM audio is the modulating signal for this system. The block diagram of the signal processing is illustrated in Fig. 2-3. SNR is related to the carrier-to-noise ratio (CNR) by the following equation:

$$\text{SNR} = \text{CNR} \cdot \text{FMI} \cdot \text{PPF} \cdot \text{E} \cdot \text{NWF} \qquad (2\text{-}1)$$

where CNR: rms carrier power to rms noise power ratio in the IF band. There is a restriction on CNR called FM threshold: CNR 10>dB Taking 1 dB margin we have CNR = 11.0 dB for the worst case.

FMI: FM improvement factor

PPF: conversion factor from rms definition to peak to peak definition, 9 dB

E: improvement factor of unweighted SNR due to emphasis, which is up to about 3 dB for color.

NWF: noise weighting factor, which accounts for the measurable improvement in SNR due to the inherent filtering of both the picture tube response and the human eye response to noise. According to CCIR the flat noise weighting factor (for AM) for monochrome TV signals is 6.1 dB and the triangular noise weighting factor (for FM without emphasis) for monochrome TV signals is 10.2 dB [14]. According to the Bell system, the latter value is 9.6 dB [7]. When the emphasis is applied, the noise is no longer triangular but is close to flat. Hence, it is reasonable to say that E · NWF is constant and independent of the emphasis. Furthermore, it can be concluded from a study of American Telephone and Telegraph (AT&T) TV link transmission performance objectives and subjective results on both color and monochrome systems that NTSC color TV signals are more susceptible to interference from flat noise than are monochrome signals. 4.0 dB instead of 6.1 dB for color was given by CCIR [14] and 5.5 dB was used by COMSAT [7]. This corresponds to 8.1 dB by CCIR and 9.6 dB by COMSAT for a triangular noise for color. The penalty of 1.5 dB is significant in satellite output power. E · NWF = 9.5 dB is used in the analysis.

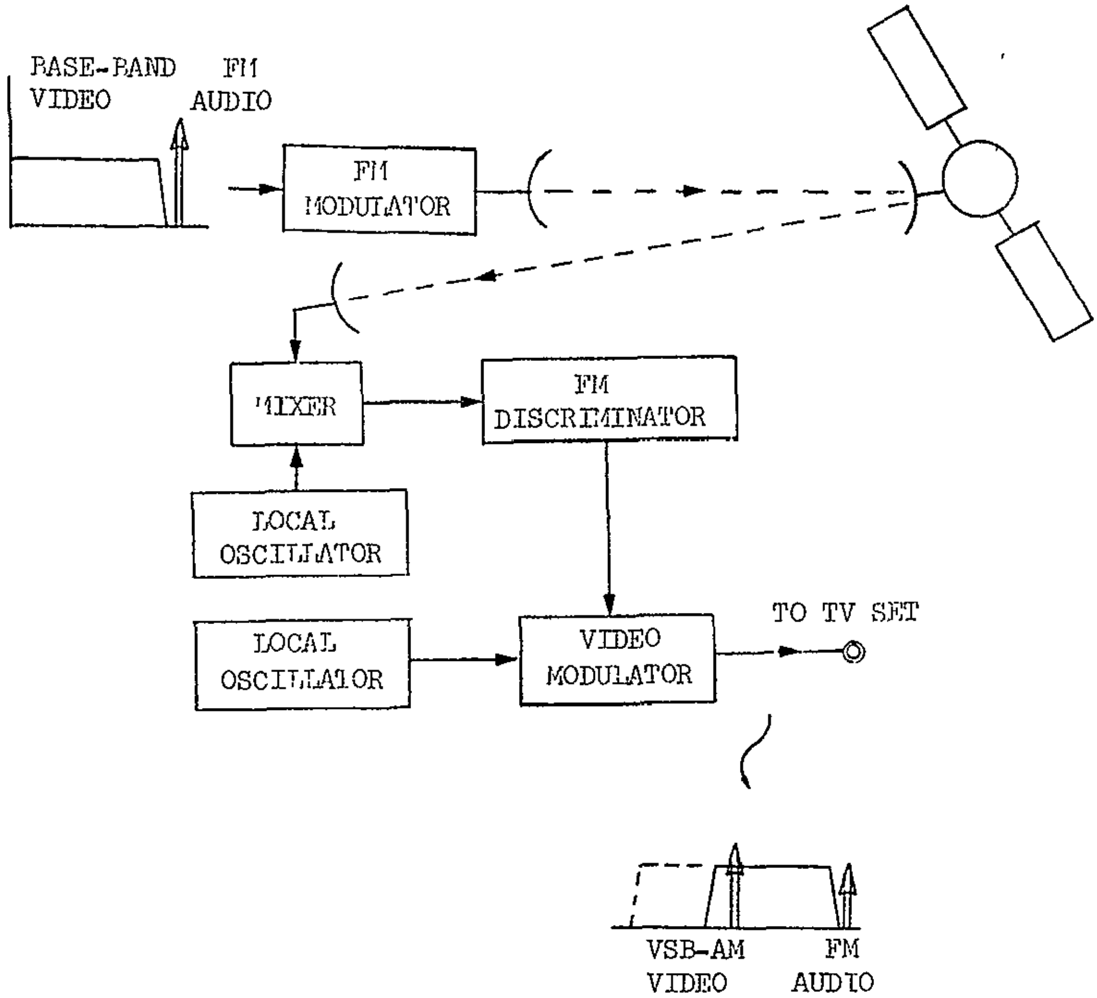


Figure 2-3 Block Diagram of the FM System

FMI factor is calculated from [15]

$$\text{FMI} = 3\beta^2 \cdot \frac{BW_{if}}{2f_m} \tag{2-2}$$

where β: modulation index

$= \Delta f/f_m$

Δf: peak frequency deviation which is determined by maximum signal level

f_m: maximum modulating frequency, 4.5 MHz

BW_{if}: the IF bandwidth 37.5 MHz, which is wider than the rf bandwidth BW_{rf}

By Carson's rule, the rf signal bandwidth BW_{rf} is related to β as [15].

$$BW_{rf} = 2(\beta+1)f_m \tag{2-3}$$

Throughout the analysis we assume 25.0 MHz rf signal bandwidth and 10% guardband which gives 27.5 MHz transmission bandwidth per channel. From Eqs. (2-2) and (2-3), the wider the rf bandwidth is, the more FMI is obtained. In other words, an FM system gives better SNR but at the sacrifice of the rf signal bandwidth.

CNR is related to the satellite EIRP as

$$\text{EIRP} = \text{CNR} \cdot L_s \cdot B_{if} \cdot K/\,(G_r/T_s) \tag{2-4}$$

where

EIRP : power radiated from the satellite referred to an isotropic antenna (effect isotropic radiated power)

L_s : total system loss including design margin

G_r/T_s : ground-terminal figure of merit. Values for model systems are given in Table 2-1.

K : Boltzman's constant = -228.6 dBW/$^\circ$K·Hz

The system loss L_s is the sum of a number of factors;

$$L_s : L_{path} + L_{rain} + L_{pointing} + L_{up\text{-}link} + L_{beam} + L_{atom} + L_{pol} + M \quad (2\text{-}5)$$

where

L_{path} : path loss (free space attenuation), 193 dB for 2.6 GHz and 206 dB for 12 GHz.

L_{rain} : rain attenuation, negligible at 2.6 GHz band but noticeable at 12 GHz band. Complete discussion on the rain margin will be given in Chapter V but in this chapter we assume 3 dB.

$L_{pointing}$ antenna pointing loss due to satellite fluctuation and wind loading on the antenna, 1 dB

$L_{up\text{-}link}$: noise contribution by the up-link, 1.0 dB (assuming CNR of up-link = CNR of downlink + 6 dB)

L_{beam} : off-beam center loss, 3 dB

L_{atom} : atmospheric absorption and long-term degradation of satellite transmitter circuit, 0.4 dB

L_{pol} : polarization loss, 0.1 dB

M : design margin for unknowns, 1 dB

The system noise temperature T_s is expressed by:

$$T_s = T_a + T_r \tag{2-6}$$

where

T_a : antenna noise temperature due to the sun, city noise, rain and so on, 50^{o}K for 2.6 GHz and 100^{o}K for 12 GHz (1% values)

T_r : receiver noise temperature in ^{o}K measured at the input of the receiver which is related to the receiver system noise figure NF_r as

$$NF_r = 10 \log_{10} \left(\frac{T_r}{290} + 1 \right)$$

In Eqs. (2-5) and (2-6), L_{rain}, $L_{pointing}$, L_{atom}, and T_a are statistical values. As the broadcasting is desired to guarantee the overall service quality for not less than 99% of the service time (uncontrolled variable 3), each factor should guarantee the quality for not less than 99% of the service time. Hence, 1% value is to be used. Finally, the design margin for unknowns M of 1 dB is added, which, however, may not be needed if a long-term experiment proves that it can be eliminated Chapter V discusses the effects of these margins and presents the probabilistic results of actually obtained signal quality.

Table 2-2 indicates the FM system parameters with the model ground-terminals shown in Table 2-1. The required EIRP is 56.2 dBW for the 2.6 GHz band which is 7 dB less than the maximum limit and 61.0 dBW for the 12 GHz band.

Table 2-2

FM Systems Parameters

	2.6 GHz	12.0 GHz
SNR	45 dB for home reception 54 dB for community reception	45 dB for home reception 54 dB for community reception
CNR	11.0 dB for home reception 20.0 dB for community reception	11.0 dB for home reception 20.0 dB for community recpetion
PPF	9.0 dB	9.0 dB
E·NWF	9.5 dB	9.5 dB
FMI	15.5 dB	15.5 dB
BW_{rf}	25.0 MHz	25.0 MHz
	1.8	1.8
f	4.5 MHz	4.5 MHz
f_{max}	13.5 MHz	13.5 MHz
G_r/T_s	0.4 dB for home reception 9.4 dB for community reception	11.0 dB for home reception 20.0 dB for community reception
BW_{if}	75.7 dB	75.7 dB
K	-228.6 dBW/^{o}K·Hz	-228.6 dBW/^{o}K·Hz
L_{path}	192.3 dB	205.6 dB
L_{rain}	0.0 dB	3.0 dB
$L_{pointing}$	1.0 dB	1.0 dB
$L_{up-link}$	1.0 dB	1.0 dB
L_{beam}	3.0 dB	3.0 dB
L_{atom}	0.4 dB	0.4 dB
L_{pol}	0.1 dB	0.1 dB
M	1.0 dB	1.0 dB
L_s	198.8 dB	215.1 dB
EIRP	56.5 dBW	61.6 dB

[B] AM System

In this system the normal broadcast TV signal, a VSB-AM video and a FM audio is used. The block diagram of the signal processing is illustrated in Fig. 2-4. The receiver is just a frequency converter. The SNR is related to CNR by

$$SNR = CNR \cdot AMI \cdot PPF \cdot NWF \qquad (2\text{-}7)$$

instead of Eq. (2-1). Here, AMI: conversion factor from CNR to rms picture power to rms noise ratio and is expressed by AMI = $(0.7)^2 \times m^2$ where m: amplitude modulation index 40~50%. 0.7 is the ratio of the white level to the synchronous pulse level. When m = 45% AMI = -10.0 dB. Equations (2-4) through (2-6) hold for the AM system, but the IF bandwidth is 6 MHz and NWF for AM is 5.5 dB as was discussed in the preceeding section. Table 2-3 indicates the AM system parameters for the same model ground terminals. The required EIRP is 78.8 dBW for the 2.6 GHz band which is 17 dB over the limitation, and 83.8 dBW for the 12 GHz band. This shows that the AM system is not feasible for satellite broadcasting from a realization standpoint. It should be noted that the amount of rf power needed on a spacecraft for such an AM system is somewhat beyond the present state of the art.

[C] AM-FM Systems

In this case, the modulating signal for the main FM carrier is a composite signal of a VSB-AM video signal and an audio AM signal. The block diagram of the signal processing is illustrated

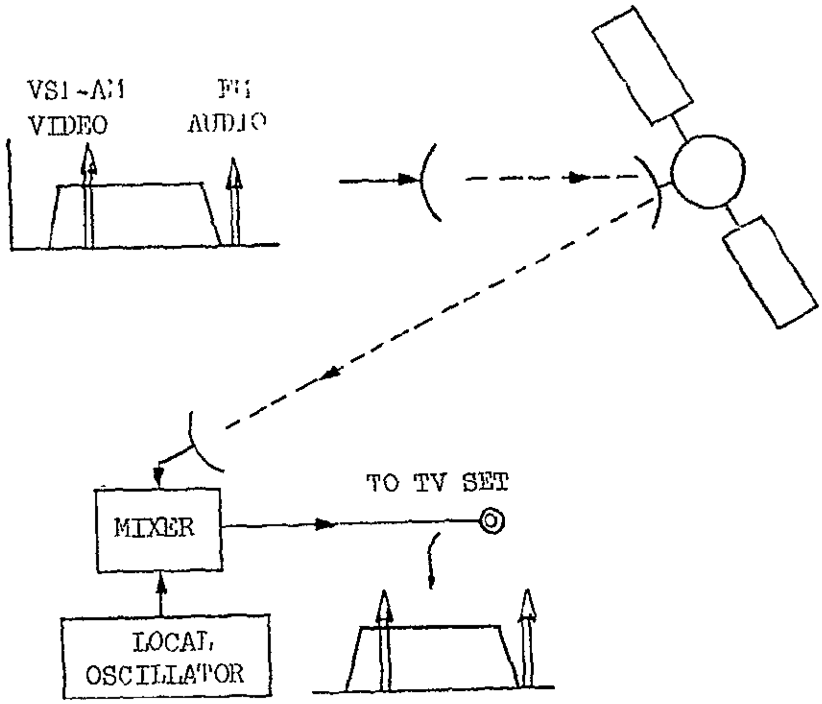


Figure 2-4 Block Diagram of the AM System

Table 2-3

AM Systems Parameters

	2.6 GHz	12.0 GHz
SNR	45 dB for direct reception 54 dB for community reception	45 dB for direct reception 54 dB for community reception
AMI PPF NWF	-10.0 dB 9.0 dB 5.5 dB	-10.0 dB 9.0 dB 5.5 dB
CNR	40.5 dB for home reception 49.5 dB for community reception	40.5 dB for home reception 49.5 dB for community reception
L_s	198.8 dB	215.1 dB
BW_{if}	6 MHz (67.8 dB)	6 MHz (67.8 dB)
K	-228.6 dBW/°K·Hz	-22.6 dBW/°K·Hz
G_r/T_s	0.4 dB for home reception 9.4 dB for community reception	11.6 dB for home reception 20.6 dB for community reception
EIRP	78.1 dBW	83.2 dBW

in Fig. 2-5. The merit of this system is in that no remodulator is needed in the receiver which makes the receiver simpler and lower in cost. The disadvantage is in that it has less FMI and NWF than the FM system, as is analyzed below.

Suppose that CNR is larger than the threshold level (10 dB) and the noise in the IF band is white. When the noise passes the FM discriminator, the noise power spectrum becomes proportional to the square of the noise frequency deviation f_n and is given by [15]

$$W_n(f_n) = A \cdot f_n^2 / (CNR \cdot BW_{if})$$

where A is a physical constant determined by the FM discriminator. This is due to the so-called triangular noise. The total noise power N_{rms} at the output of the FM discriminator is given by an integration, resulting in

$$N_{rms} = \frac{A}{3} \cdot \frac{f_h^3 - f_l^3}{CNR \cdot BW_{if}}$$

where

f_h : highest noise frequency in the band-pass filter before the FM discriminator

f_l : lowest noise frequency in the band-pass filter before the FM discriminator

The rms modulating signal power output $S_{AM\ signal}$ after the FM discriminator is given by

$$S_{AM\ signal} = \frac{A}{2} \quad (\Delta f)^2$$

where Δf is the maximum frequency deviation. Therefore,

$$SNR_{AM\ signal} = S_{AM\ signal}/N_{rms} = \frac{3}{2} \cdot \frac{\Delta f^2}{f_h^3 - f_l^3}$$

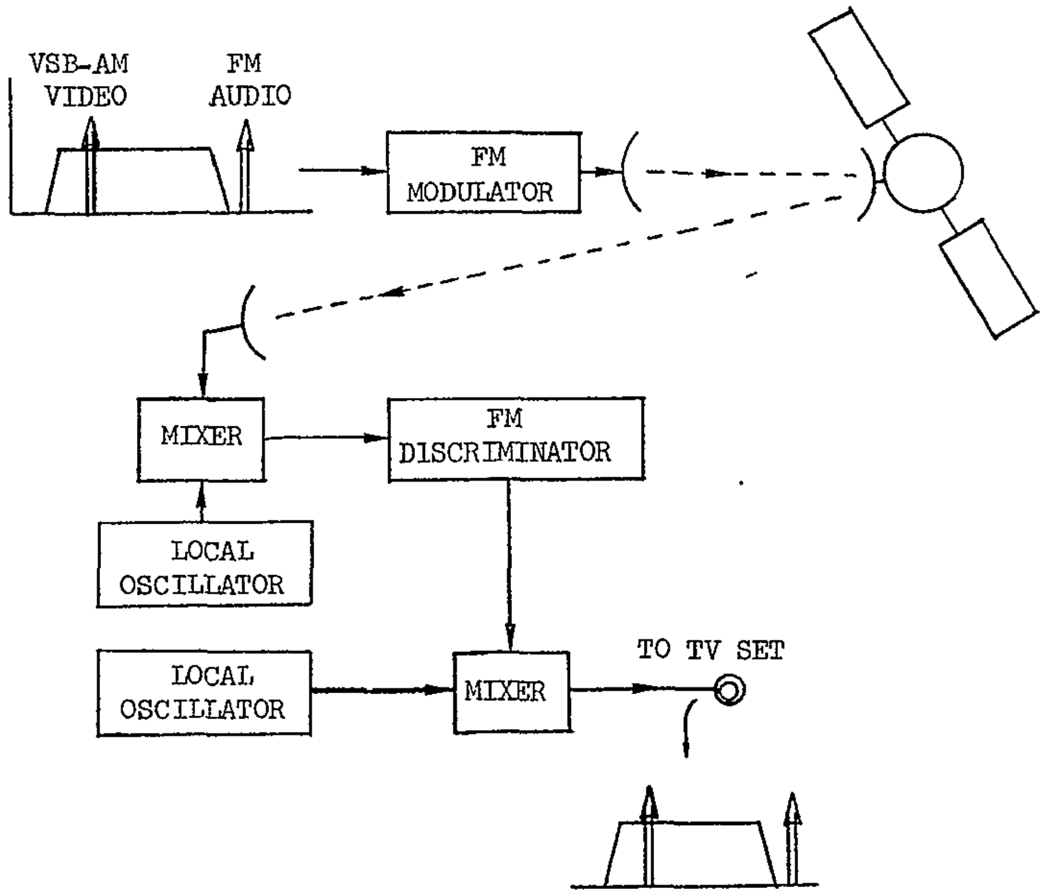


Figure 2-5 Block Diagram of the AM-FM System

and

$$FMI_{AM\ signal} = \frac{3}{2} \cdot \frac{\Delta f^2}{f_h^3 - f_l^3} \cdot BW_{if} \qquad (2\text{-}8)$$

When $f_l = 0$ and $f_h = f_m$ (the FM system), Eq. (2-8) coincides with Eq. (2-2).

Now a correction factor for Eq. (2-8) is needed because the video signal occupies only some portion of the AM signal. Figure 2-6 shows the waveform of the modulating signal (VSB AM signal) where the video signal is assumed to be a sinusoidal wave for simplicity. The P-P video voltage is m times the P-P subcarrier voltage and (m/1+m) times the P-P modulating signal voltage. As the frequency deviation is in proportion to the P-P maximum voltage, the rms video signal power to rms noise power ratio SNR_{video} is $\left(\frac{m}{1+m}\right)^2$ time $SNR_{AM\ signal}$. Hence, finally

$$FMI = FMI_{AM\ signal} \cdot \left(\frac{m}{1+m}\right)^2$$

$$= \frac{3}{2} \cdot \frac{\Delta f^2}{f_h^3 - f_l^3} \cdot \left(\frac{m}{1+m}\right)^2 \cdot BW_{if} \qquad (2\text{-}9)$$

and the rf bandwidth of the AM-FM system is given by [16]

$$BW_{rf} = 2(\Delta f + f_v)$$

where f_v: sub-carrier frequency of modulating signal $= f_h - 4.5$ MHz $= f_l + 1.25$ MHz.

m : modulation index for AM, 40~50%. The modulation index in this case is $\Delta f/f_h$. Figure 2-7 shows FMI for the AM-FM systems for various $f_l (m = 45\%)$.

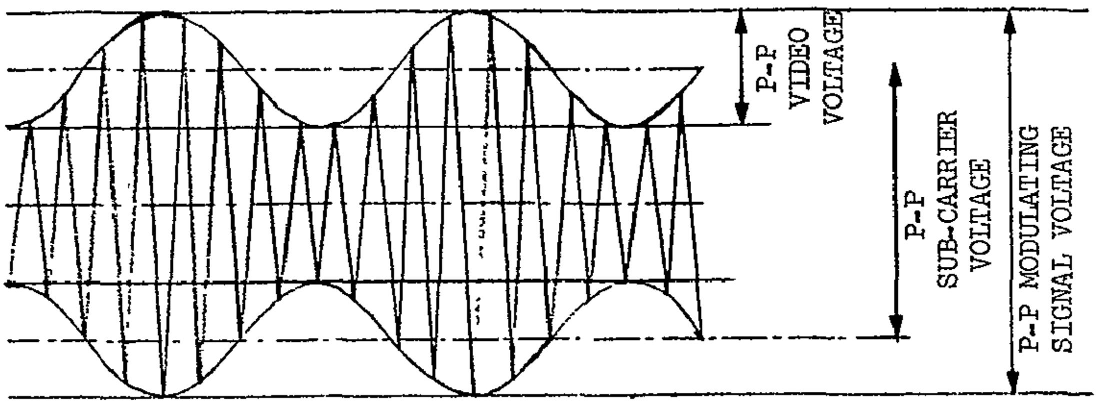
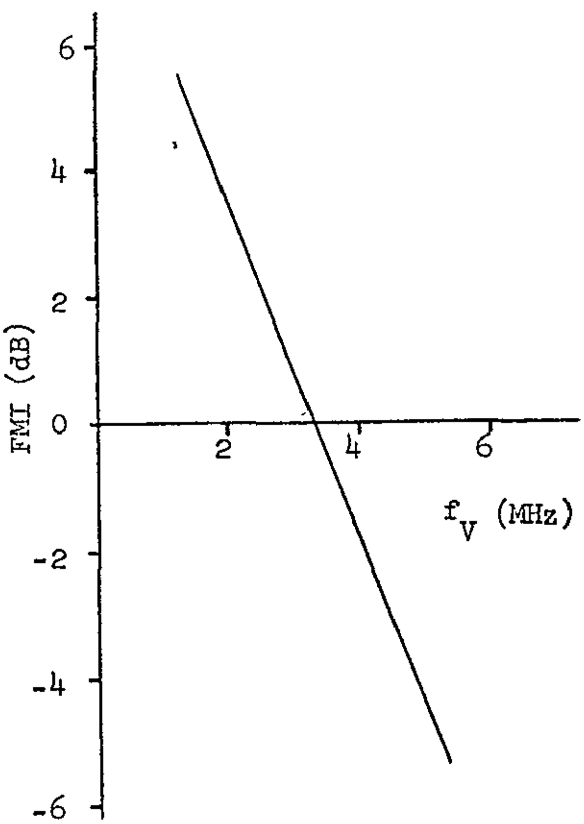


Figure 2-6 Waveform of the VSB-AM Video Signal
(modulating signal of the AM-FM system)



(BW_{rf} = 25.0 MHz and FMI for FM systems with the same bandwidth = 15.5 dB)

Figure 2-7 FMI for the AM-FM System

NWF depends on the noise spectrum as was described in Section [A]. In case of AM-FM, the noise spectrum becomes closer to flat noise with the increase in f_1. Unfortunately, there is no report on this experiment therefore the worst case of 5.5 dB for NWF is taken for all the cases of AM-FM systems for safety.

Table 2-4 shows the system parameter for the same model ground terminals as the foregoing two cases. The difference from the FM system comes from the differences in FMI and NWF, which sums up to 9.5 dB for f_1 = 1.25 MHz. The required EIRP for the 2.6 GHz band is 70.2 dBW which is 6.8 dBW over the limitation. Therefore, the AM-FM system with MHz rf bandwidth is not available for this frequency band. EIRP for the 12 GHz band is 75.0 dBW and, at present, practically unavailable although there is no power limitation in this frequency band.

EIRP for the AM-FM system can be reduced at the sacrifice of the rf signal bandwidth. As will be shown in Appendix D, the same EIRP as the FM system is available at twice as wide bandwidth as for the AM-FM system.

2-2-3 Decision of Modulation System

It may therefore be concluded that only the FM system can afford the direct reception. 61.0 dBW of EIRP for the 12.0 GHz band is achieved by the 350W transmitter and 2.4' antenna which fully covers one third of the U.S.A. Further sensitivity study will be done in Chapter V.

Table 2-4

AM-FM Systems Parameters

	2.6 GHz	12 GHz
SNR	45 dB for home reception 54 dB for community reception	45 dB for home reception 54 dB for community reception
CNR	25.0 dB for home reception 20.0 dB for community reception	25.0 dB for home reception 20.0 dB for community reception
PPF E·NWF FMI BW_{rf} f_h f_v f_{max}	9.0 dB 5.5 dB 5.5 dB 25.0 MHz 1.95 5.75 MHz 1.25 MHz 11.25 MHz	9.0 dB 5.5 dB 5.5 dB 25.0 MHz 1.95 5.75 MHz 1.25 MHz 11.25 MHz
G_r/T_s	0.4 dB for home reception 9.4 dB for community reception	11.6 dB for home reception 20.6 dB for community reception
BW_{if}	37.5 MHz	37.5 MHz
K	-228.6 dBW/°K·Hz	-228.6 dBW/°K·Hz
L_s	198.8 dB	215.1 dB
EIRP	70.5 dBW	75.6 dBW

CHAPTER III

DECISION ANALYSIS FOR OPTIMIZATION IN RECEIVER DESIGN

3.1 DESCRIPTION OF GROUND-TERMINAL

In this chapter, the application of the decision analysis model described in Chapter II for the optimization in the design of the 12 GHz receiver where state of the art technologies have to be converted to mass-producible technologies is carried out.

The ground-terminal consists of a parabolic antenna and a receiver, which consists of an outdoor unit and an indoor unit. The receiver configuration alternatives and decision analyses are discussed in Appendix B-1. The outdoor unit is placed at the focus of the parabola by three arms. The input signal is a circularly polarized (CP) FM TV signal with 25 MHz bandwidth which is broadcast from a satellite. Polarization alternatives are discussed in Appendix B-2. The analysis will be focussed on the outdoor unit design which is the key to the whole ground-terminal design. Figure 3-1 shows the block diagram of the outdoor unit.

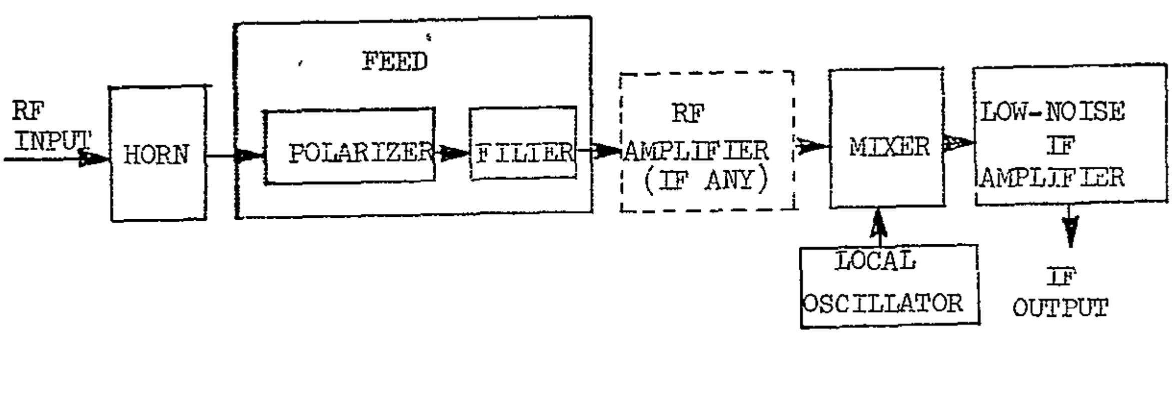


Figure 3-1 Block Diagram of Outdoor Unit

3.2 DECISION MODEL

The decision analyses for subsystems such as the local oscillator device, mixer and polarization are in Appendix B-2. The decision model discussed in Chapter II is applied. The uncontrolled variables, decision variables and outcomes for the optimization process of this section are listed below.

Uncontrolled variables (design constraints):

Noise figure (NF); 10 dB or less

LO frequency stability; $\pm$ 3 MHz or less for $\pm$ 50°C temperature change, 0~100% humidity change, up to 800 mmHg air pressure change and $\pm$ 10% bias voltage change

Polarization, Circular Polarization (CP)

Input signal; 12.0 GHz, $(\text{baseband}_{\text{video}} + \text{FM}_{\text{audio}})$-FM
25 MHz rf signal bandwidth

Spurious radiation; within FCC limit

Image rejection rate, -15 dB or less

Power supply; 12 V DC

Output; fed to indoor unit through a coaxial cable

Decision variables:

Subsystems; detailed discussions are found in Appendix B. The results are summarized in Table 3-1.

Intermediate frequency; f_{IF}

Outcomes:

Cost, parts
manufacturing
installation
overhead and profit

Table 3-1

Decision Variables

Subsystems	Waveguide	Microstrip	Stripline
Horn	Circular waveguide	Printed	Not available
Feed	Circular or square waveguide	Printed	NA
RF amplifier (if any)	Circulator + dummy load + waveguide TDA	Printed circulator, dummy load & TDA	NA
Polarizer	Integral with feed	Integral with feed	NA
Preselector	Integrate with feed (high-pass or band-pass)	Edge-coupled image	Edge-coupled image, reject or band-pass
Mixer	Single-ended (SE) or singly-balanced (SB) orthomode	SE or SB hybrid	SE or SB hybrid
LO filter (necessary only if SE is used)	Inductive post filter	Edge-coupled	Edge-coupled
Local oscillator (LO)	Gunn oscillator with invar stabilizing cavity	MIC Gunn oscillator coupled to quartz stabilizing cavity	NA
Intermediate frequency	Any possible	Any possible up to 1.5 GHz	NA

Outcomes (cont.)

Performances; gain
noise figure
IF frequency stability
life time
reproducibility and repairability in the field

Capability of future modifications;

pre-amplifier for better NF
polarization
modification to multi-channel transmission

3.3 INTERACTION MODEL

Under the constraints of the stated variables we can derive a number of combinations of subsystems from the decision variables listed in Table 3-1. Four major alternatives were selected and analyzed.

[A] The waveguide system

A circular waveguide horn, circular to rectangular transition, rectangular waveguide, orthomode mixer and rectangular waveguide Gunn oscillator

[B] The Waveguide-Microwave Integrated Circuit (MIC) hybrid system (1)

A circular waveguide horn and feed, MIC mixer and rectangular waveguide Gunn oscillator

[C] Waveguide-MIC hybrid system (2)

A circular waveguide horn and feed, MIC Gunn oscillator and MIC mixer

[D] MIC system

A horn, tunnel diode amplifier circuit (TDA), mixer and Gunn oscillator. All are made using MIC techniques.

In the case of [A] and [B], the Gunn oscillator is stabilized by an invar stabilizing cavity. On the other hand, a polycrystalline quartz cavity is used for stabilization in the case of [C] and [D].

3.3.1 Performance Comparison

3.3.1.1 Integrability

[A] Waveguide system

The whole packaging is die cast in two identical pieces. Figure 3-2 shows the overview. The invar cavity which can be cast[17] or machined from tubular stock[18] is attached to the Gunn oscillator. The first IF board (low noise IF amplifiers) is packed in the packaging. A two-pole bandpass filter is used to avoid extra length of the packaging.

[B] Waveguide-MIC hybrid system (1)

The whole packaging is die cast in two identical pieces and the cast invar cavity is attached and electrically coupled to the Gunn oscillator as was the case with [A]. A mixer and a TDA circuit, if any, are integrated on an aluminum substrate. Figures 3-3 and 3-4 show the overview of the packaging and the sketch of the cast aluminum piece. A high-pass filter is used which needs no adjustment.

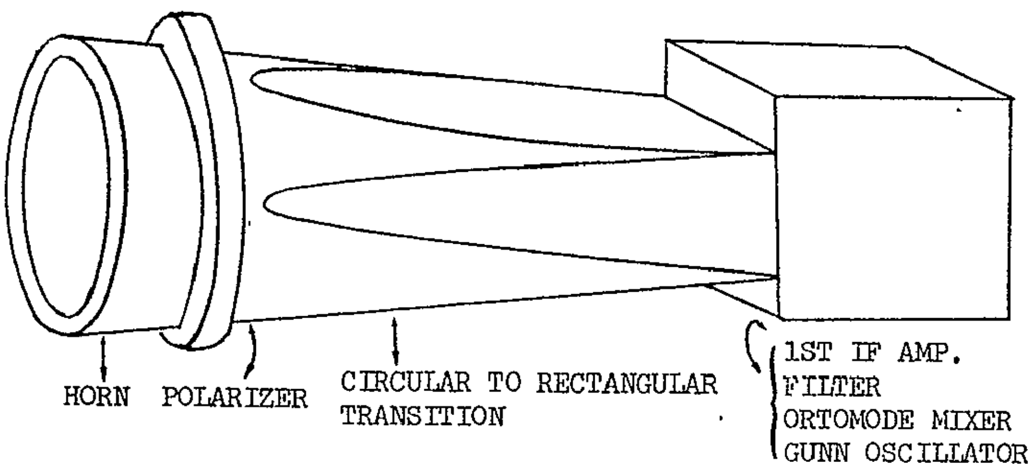


Figure 3-2 Sketch of Waveguide System (alternative [A])

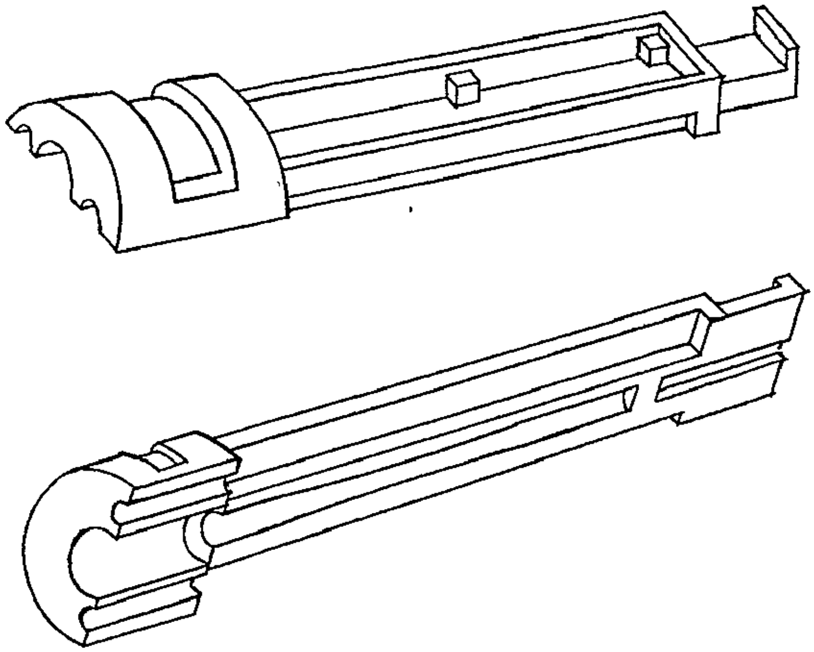


Figure 3-3 Sketch of Cast Packaging (alternative [B])

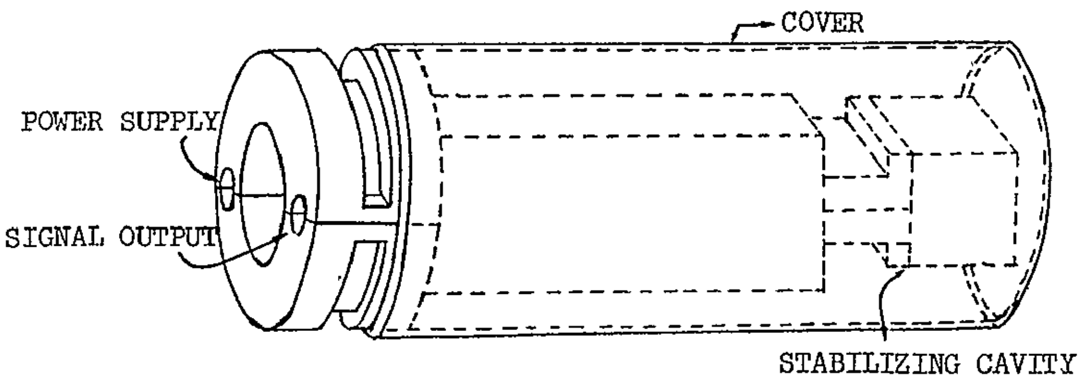


Figure 3-4 Sketch of Assembled Packaging (alternative [B])

[C] Waveguide-MIC hybrid system (2)

The system differs from [B] in that the Gunn oscillator is made by MIC and stabilized by a quartz stabilizing cavity. Figure 3-5 shows the sketch.

[D] MIC system

The feed is printed on an aluminum substrate using two crossed dipoles. The detected signal is fed to a TDA through two coaxial cables and then fed to a MIC mixer. The Gunn oscillator is the same as for [C], whose output power is fed to the MIC mixer Figure 3-6 shows a sketch of it. The packaging is die cast.

3.3.1.2 Assembly and installation in the field

The packaging has to be mechanically strong enough to be fixed at the focus of the parabola by three arms in the proper direction toward the satellite. The rotation around the axis does not matter as long as circular polarization is used. A cylindrical package is desirable for the installation. The axial length of the package is longer in the order of [A], [B], [C], and [D]. [A] may need an extra support, but there will not be much difference in the assembly cost among them.

3.3.1.3 Electrical performances

The MIC feed of alternative [D] is lossiest, giving about 2 to 3 dB worse noise figure than the other three alternatives, while there is not much difference among the others. Hence, the alternative [D] requires a TDA circuit to get the equivalent system noise figure to the others.

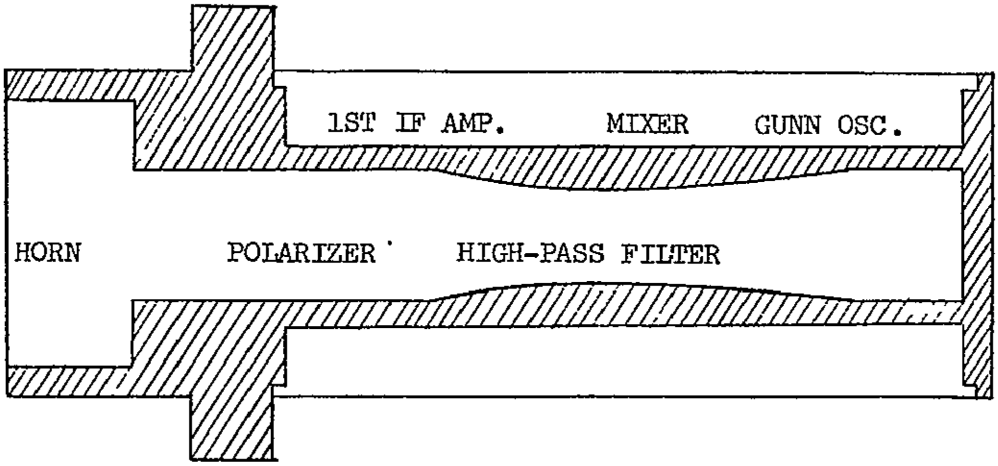


Figure 3-5 Cross-sectional View of Waveguide-MIC System (2) (alternative [C])

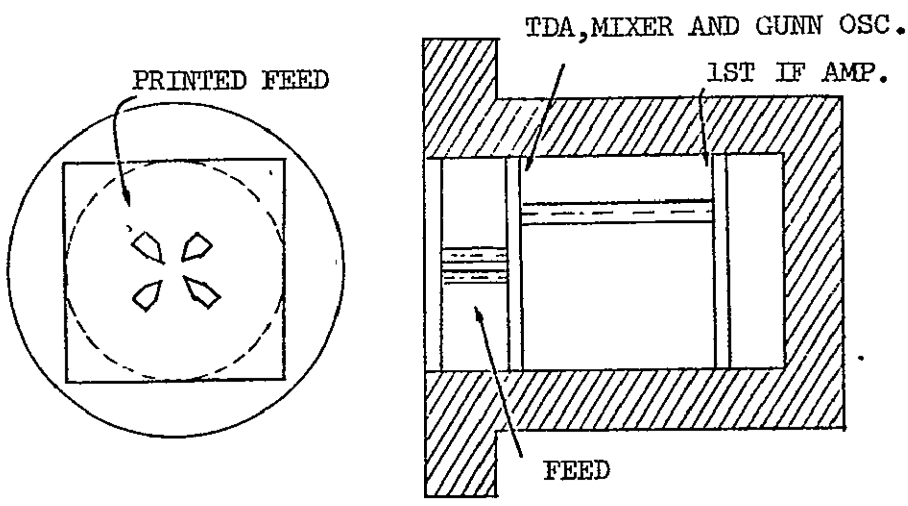


Figure 3-6 Sketch of MIC System (alternative [D])

The IF has to be greater than 120 MHz to get -15 dB image rejection. On the other hand, the cost of an IF amplifier circuit increases with the increase in f_{IF} if the same NF is assumed. Hence, 120 MHz is the optimum frequency where the circuit technology is well developed in TV industries.

3.3.2 Cost Comparison

Tables 3-2, 3-3, 3-4 and 3-5 show the estimated cost of the alternatives [A], [B], [C] and [D] respectively, where common items are omitted just for the comparison Figure 3-7 shows the differential costs of these four alternatives for various production quantities up to 10^5 units per year. Cost quotations for more than 10^5 units per year were not prepared because it is highly unlikely that one manufacturer would ever be asked to produce this quantity. If there were a sufficient demand, more than one supplier would be involved and the economies of scale would not continue. The cost information used here came mainly from quotes by manufacturers and discussion with production engineers in the industry.

3.3.3 System Diversity

3.3.3.1 The addition of a preamplifier circuit

There is a good possibility that a preamplifier will be required to get a better noise figure in the future. A tunnel diode amplifier (TDA) would be the best alternative to get a 7 dB system noise figure at 12 GHz because of its lower cost per performance than other possible alternatives. These other alternatives include the para-

Table 3-2

Cost of Alternative [A] (Waveguide System)

(Common items are omitted in the table)

Items	Quantity	Production Quantity (units/year)				
		10	10^2	10^3	10^4	10^5
[Packaging]						
diecast aluminum (150% of B)	2	630	66.6	11.1	5.7	5.5
[Waveguide Mixer]						
1. diodes (HP 5082-2235)	2	29	22	19.5	140	12.0
2. assembly & adjustment including a screw cost		5	3	2	1	0.8
3. testing		3.7	2.7	2.2	1.7	1.4
(sub-total)		(37.7)	(27.7)	(21.7)	(16.7)	(14.2)
[Waveguide Gunn Oscillator]						
1. diode (Fairchild GD (x) 510C)	1	26	19	14	12.6	12.0
2. assembly, adjustment & shielding		8	4.5	3	1.5	1.0
3. invar cavity material		18	10	2.5	1.0	0.8
4. machining		5	2	1.5	1.0	1.0
5. testing		6.3	3.9	2.2	1.8	1.6
(sub-total)		(63.3)	(39.4)	(23.2)	(17.9)	(16.4)
[Final Assembly & Testing]						
(20% of total cost)		183	33.6	1.4	10.1	8.9
Total Factory Cost		\$914	\$168	\$70.0	\$50.5	\$44.5

Table 3-3

Cost of Alternative [B] (Hybrid Systems (1))

(Common items are omitted in the table)

Items	Quantity	Production Quantity (units/year)				
		10	10^2	10^3	10^4	10^5
[Packaging]						
1. diecast aluminum	2	380	40	6.6	3.4	3.4
2. testing		43	4 5	0.8	0.4	0.3
(sub-total)		(423)	(44.5)	(7.4)	(3.8)	(3.7)
[MIC Mixer]						
1. chips (HP-5082-0023)	2	9.0	7.0	3.5	2.5	2.0
2. aluminum substrate	1/2 x 3/4 in.	.8	.6	.5	.4	.3
3. mask		5.0	.5	0	0	0
4. etching & cleaning		4	3	2	1.6	1.5
5. bonding & die attaching		5	3.5	2.5	2.0	1.8
6. testing		5	3.5	2.5	2 0	1.8
7. shielding		4.0	2 0	1.0	0.8	0.7
8. transitions		5.0	3.5	2.5	2.0	1.8
9. final testing		4.2	2.6	1.6	1.3	1.1
(sub-total)		(42.0)	(26.2)	(16.1)	(12.6)	(11.0)
[Waveguide Gunn Oscillator] same as [A]		63.3	39.4	23 3	17.9	16.4
[Final Assembly & Testing] (20% of total cost)		132	27 6	11.7	8.6	8.0
Total Factory Cost		$660	$138	$58.5	$43	$40

Table 3-4

Cost of Alternative [C] (Hybrid Systems (2))

(Common items are omitted in the table)

Items	Quantity	Production Quantity (units/year)				
		10	10^2	10^3	10^4	10^5
[Packaging]						
diecast aluminum (80% of [B])	2	338	35.5	5.9	3.0	2.6
[MIC Mixer]						
same as [B]		42.0	26.2	16.1	12.6	11.0
[MIC Gunn Oscillator]						
1. chips, Fairchild GD (x) 510C	1	18	10	7	5	4.5
2. risk (20%)		3.6	2	1.4	1.0	0.9
3. heat sink (machined & attached)	1	5	4	3	2.5	2.0
4. etching & cleaning		4	3	2	1.6	1.5
5. bonding & die attaching		5	3.5	2.5	2.0	1.8
6. aluminum substrate		.8	.6	.5	.4	.3
7. mask		5	.5	0	0	0
8. quartz	3/4 x 2/3 in.	10	6	4	3	2.8
9. packaging & sealing		5	4	33	2.5	2.0
10. testing (RF property)		5	3.5	2.5	2	1.8
11. final testing		6.8	4.4	2.9	2.2	1.9
(sub-total)		(68.2)	(44.5)	(29.3)	(22.2)	(19.5)
[Final Assembly & Testing]		112	26.5	12.8	9.2	8.0
Total Factory Cost		$560	$133	$64	$47	$41

Table 3-5

Cost of Alternative [D] (MIC System)

(Common items are omitted in the table)

Items	Quantity	Production Quantity (units/year) 10	10^2	10^3	10^4	10^5
[Packaging]						
diecast aluminum (50% of [B])	2	212	22	3.7	1.9	1.7
[MIC Mixer]						
same as [B]		42.0	26.2	16.1	12.6	11.0
[MIC Feed]						
1. substrate	2 x 2"	8	6	5	4	3
2. mask		5.0	0.5	0	0	0
3. etching & cleaning		4	3	2	1.0	1.5
4. coaxial feed		2.0	1.5	1.0	8	.7
5. testing		1.5	1.2	0.9	0.7	0.6
(sub-total)		(15.5)	(12.1)	(8.9)	(7.1)	(5.8)
[MIC PDA]						
1. parts and assembly including a circulator circuit		270	170	100	60	40
2. testing		30	20	11	7	4.5
(sub-total)		(300)	(190)	(111)	(67)	(44.5)
[MIC Gunn Oscillator] same as [C]		68.2	44.5	29.3	22.2	19.5
[Final Assembly & Testing] (20% of parts cost)		162	75	42	27	20
Total Factory Cost		$800	$370	$211	$138	$103

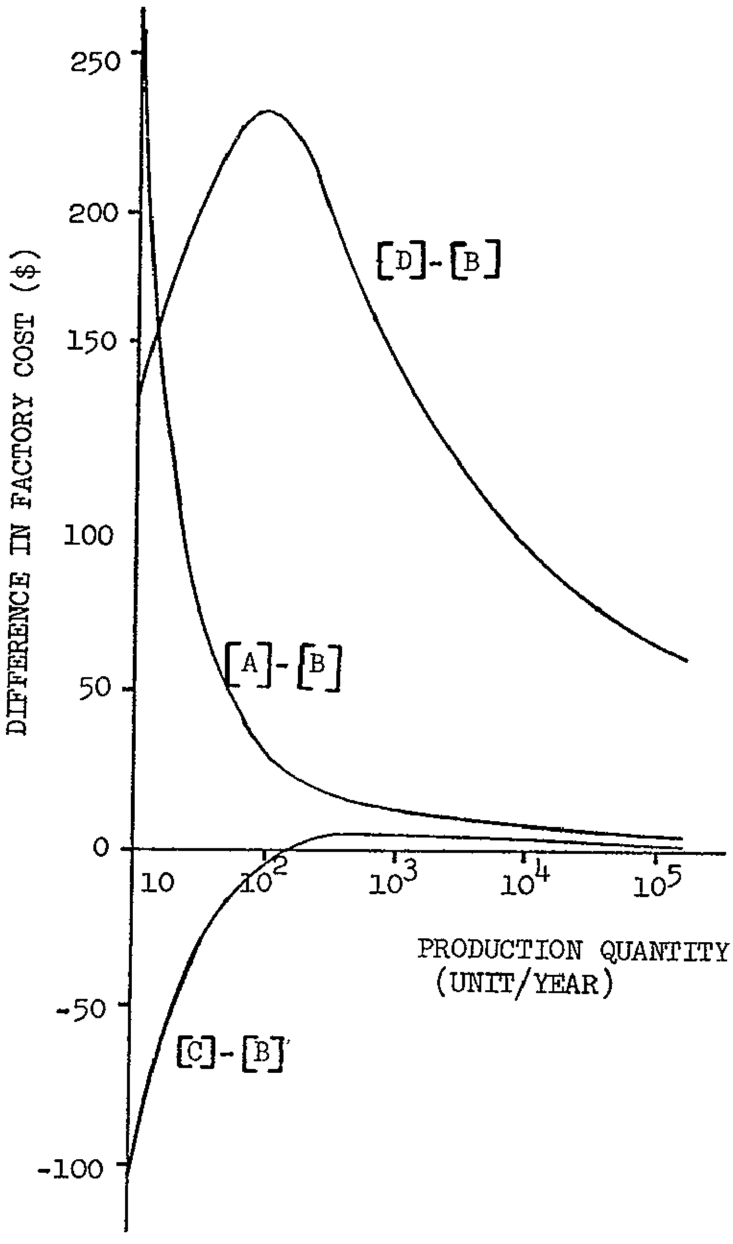


Figure 3-7 The Difference in Cost of Alternative [A], [C] and [D] from Alternative [B]

metric amplifier which gives better performance but at a much higher cost and the Gunn effect amplifier which has a higher noise figure. The TDA circuit consists of a tunnel diode amplifier, circulator and dummy load Both a waveguide circuit and a MIC circuit are available. An uncooled parametric amplifier (UPA) is the best alternative to get about 4 dB or less system noise figure

The alternative [A] is hard to modify for the additional TDA or UPA circuit, and an entirely new packaging is required, since the old packaging is useless. On the other hand, the alternatives [B] and [C] are easy to modify, where an additional MIC TDA or UPA circuit is installed between the filter and the mixer. The alternative [D] has poorer potential for better NF because it already has a TDA.

The additional cost of a TDA amplifier circuit for the systems [B] and [C] is the same as the cost of a TDA for the system [D] which is shown in Table 3-5.

3.3.3.2 Polarization

Linear polarization (LP) might be used instead of circular polarization. In that case, we have only to pull out a polarizer from the antenna unit, in the cases of [A], [B] and [C]. The alternative [D] is the same for CP and LP. Since a polarizer does not cost much, it is important to design the horn and feed system so that it works both for CP and LP.

3.3.3.3 Modification to multi-channel transmission

All the alternatives are modifiable to multi-channel transmission, which will be discussed in Appendix D, by lowering the local oscillator frequency a bit and adding a UHF tuner circuit.

3.4 DECISION MAKING

From the result of the cost analysis shown in Fig. 3-7, it was concluded that the alternative [B] or [C] should be chosen. The intermediate frequency is chosen at 120 MHz. [C] has an advantage in small quantities (below 200) and [B] is cheaper in large quantities (over 200) but only slightly.

From performance comparison, [B] is superior to [C] in frequency stability. That is to say, [B] has the stability of $\pm$ 2 MHz maximum, whereas [C] has $\pm$ 3 MHz maximum. This means that [B] has a lower noise figure than [C] by about 0.5 dB.

Other performances (capability of system modification etc.) are the same. Therefore, it is concluded that the alternative [B] is the best in quantities above 100, at least at present. However, it should be kept in mind that the alternative [C] provides a good possibility for reducing the cost below that of [B] in the future when the use of MIC technology becomes more widespread in the consumer electronics industry.

3.5 COST OF ALTERNATIVE [B]

Tables 3-6 and 3-7 show the factory cost of the outdoor unit and the indoor unit, respectively. Figure 3-8 illustrates the final results. The selling price is the factory cost multiplied by 2.5~3.0[19]. Figure 3-8 shows the case of 3.0.

Table 3-6

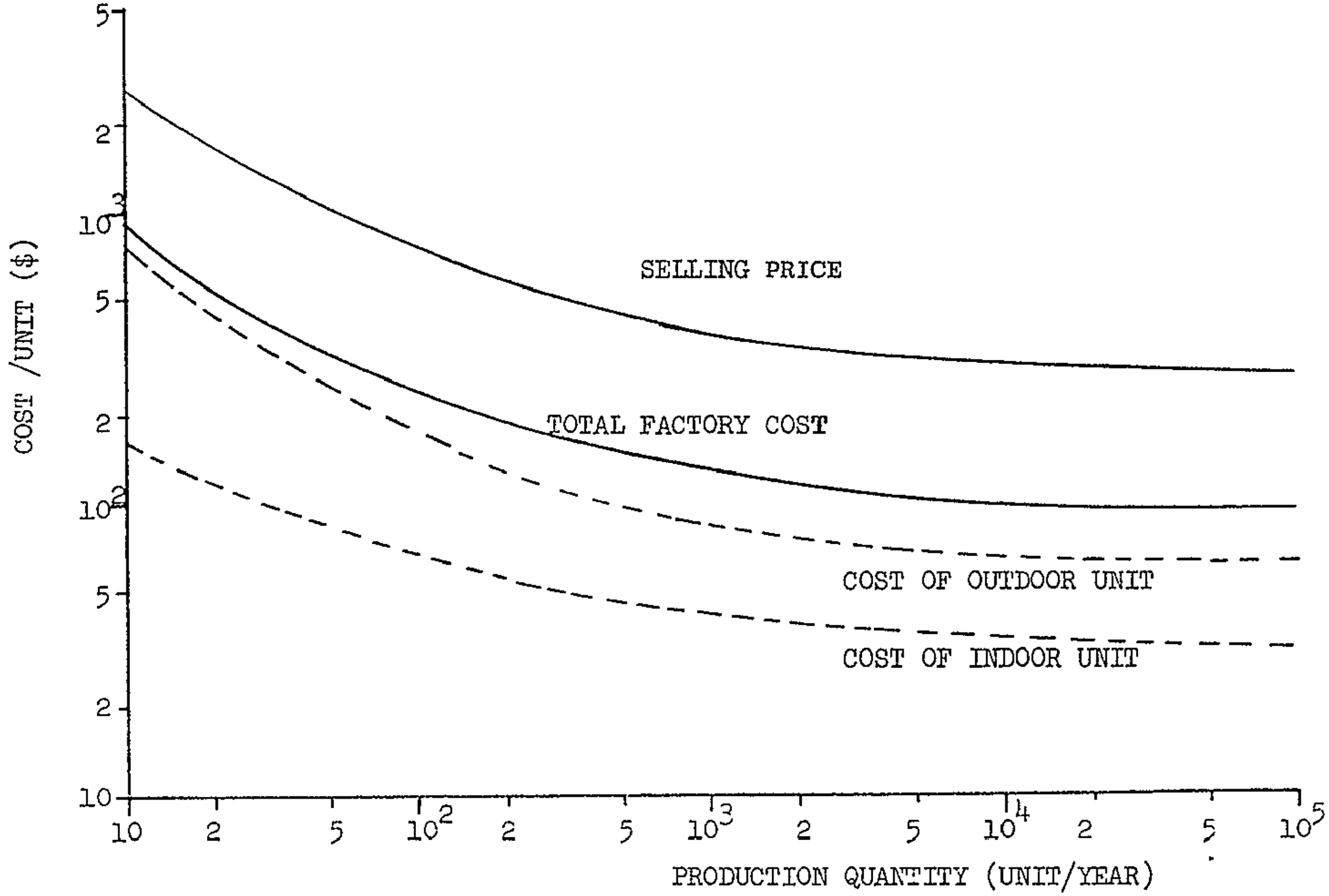
Factory Cost of an Outdoor Unit

Items	Quantity	Production Quantity (units/year)				
		10	10^2	10^3	10^4	10
[Packaging]						
1. diecast aluminum	2	380	40	6.6	3.4	3.0
2. cover	1	3	2	1	0.50	0.50
3. miscellaneous parts		2.50	1.0	0.50	0.25	0.20
4. machining		5.0	2.0	0.60	0.35	0.30
5. testing		43.5	5.0	1.0	0.5	[illegible]
(sub-total)		(435)	(50)	(10)	(5)	(4.1)
[Waveguide Gunn Oscillator]						
1. gunn diode	1	26	19	14	12.	12
2. miscellaneous		5	4.5	3	1.5	1.0
3. stabilizing cavity material		18	10	2.5	2.2	2.0
4. machining and sealing		5.0	2.0	1.5	1.0	1.0
5. testing		6.5	4.0	2.5	2.0	1.8
(sub-total)		(63.5)	(39.5)	(23.5)	(19.3)	(17.)
[MIC Mixer]						
1. chips	2	9	7	3.5	25	2.0
2. alumina substrate	1/2 x 3/4"	0.8	0.6	0.5	0.4	0.3
3. mask		5.0	0.5	0	0	0
4. etching and cleaning		4	3	2	1.6	1.5
5. bonding and die attaching		5	3.5	2.5	2.0	1.5
6. testing		5	3.5	2.5	2.0	1.5
7. shielding		4	2	1.0	0.8	0.7
8. transitions	2	5	3.5	2.5	2.0	1.8
9. final testing		4.2	2.6	1.6	1.3	1.1
(sub-total)		(42.0)	(26.2)	(16.1)	(12.6)	(11.0)
[1st IF Board]						
1. resistor	12	0.84	0.72	0.48	0.42	0.42
2. ceramic capacitor	10	0.30	0.25	0.24	0.22	0.20
3. trimmer capacitor	3	0.45	0.36	0.31	0.28	0.25
4. inductors	3	0.15	0.15	0.12	0.09	0.09
5. chokes	3	0.30	0.24	0.18	0.12	0.12
6. transistor (3N159)	1	2.18	1.58	1.32	1.25	1.15
7. transistor (2N 3563)	2	0.58	0.36	0.30	0.28	0.26
8. circuit board	8 x 8"	0.24	0.24	0.20	0.16	0.16
9. printing		5	1.25	0.75	0.62	0.55
10. voltage regulator	1	2.20	1.75	1.00	1.50	1.40
11. assembly and testing (20% of sub-total)		3.00	1.80	1.40	1.25	1.20
(sub-total)		(15.30)	(8.60)	(6.90)	(6.25)	5.75
[Wiring]						
1. cables	20m	5	5	3.5	2.5	2.5
2. connectors		2	1.5	1	1	1
3. miscellaneous		1	1	0.5	0.5	0.5
4. testing		0.9	0.9	0.6	0.5	0.5
(sub-total)		(8.9)	(8.4)	(5.6)	(4.5)	(4.5)
[Assembly and Testing] (20% of total cost)		162.0	34.0	16.5	13.0	12.0
[Compartment] (for shipping)	1	20	10	5	4	4
Total Factory Cost of Outdoor Unit		\$ 730	\$ 171	\$ 83	\$ 65	\$ 60

Table 3-7

Factory Cost of an Indoor Unit

Items	Quantity	Production Quantity (units/year)				
		10	10^2	10^3	10^4	10^5
[Parts]						
1. filter & 2nd IF amplifier		13.42	8.41	4.82	4.00	3.50
2. limiter & discriminator		9.42	6.14	4.55	4.00	3.50
3. remodulator		6.92	4.45	2.53	2.30	2.00
4. power supply		15.00	10.00	7.00	6.00	5.00
5. circuit boards		20.00	5.00	3.00	2.50	2.30
6. switch, wires, miscellaneous		2.00	1.50	1.00	.80	.70
7. housing		39.00	9.00	4.50	4.00	3.50
8. parts testing		11.8	5.0	3.0	2.6	2.3
(total parts cost)		(118.0)	(49.50)	(30.5)	(26.0)	(23.0)
Assembly (10% of total costs)		16.0	6.8	4.1	3.5	3.1
Compartment		10	5	2.5	2	2.0
Total Factory Cost of Indoor Unit		$160	$68	$41	$35	$31



(Total factory cost is the sum of an outdoor unit cost and an indoor unit cost; selling price is the total factory cost multiplied by 3.0)

Figure 3-8 Receiver Cost & Price vs Production Quantity Per Year

CHAPTER IV

GROUND-TERMINAL DESIGN

4.1 INTRODUCTION

The recent development in microwave solid-state technology and circuit technology have offered a promising future for small-size low-cost microwave receivers for communication use. This chapter describes the design of a 12 GHz mass-producible low-cost receiver system by using state-of-the-art microwave technologies. The study of frequency stability of the negative resistance diode oscillator has made a highly stabilized Gunn oscillator realizable for a local oscillator for low-cost communication applications. The combination of a newly developed low-loss high-efficiency feed system[20], consisting of a horn, polarizer and high-pass filter, and hybrid integrated circuit mixer technologies made the 8 dB system noise figure and the 15 dB image rejection rate for 120 MHz intermediate frequency possible. The ground-terminal consists of a parabolic antenna system, an outdoor unit and an indoor unit as was stated in Chapter III and is described in detail in Appendix B-1. This chapter summarizes the design procedure of each stage and measures performances of the working models built. Detailed discussions on the design are given in Appendix C.

4.2 OUTDOOR UNIT

The outdoor unit is placed at the focal point of a parabolic antenna. This configuration eliminates any signal attenuation due to a transmission line which otherwise has to be installed between the horn and the receiver, but, on the other hand, causes a severe problem on the frequency stability of a local oscillator because the effect of

environmental conditions (temperature, humidity and air pressure change) on the oscillator performance is considerably important. Therefore, a highly stabilized oscillator is needed for the outdoor unit. A Gunn diode oscillator with a reaction type stabilizing cavity analyzed in Appendix A is used. The packaging which composes a feed system (horn, polarizer, and high-pass filter), local oscillator main cavity and housing for a mixer and the first stage IF amplifier is to be made by two identical aluminum diecast pieces. The mixer is made by a hybrid integrated circuit. The local oscillator consists of a main cavity with a Gunn diode mounted in it and a reaction type invar stabilizing cavity. The first IF amplifier consists of a low-noise field-effect-transistor (FET) amplifier stage followed by three stage bipolar transistor amplifiers. Detailed discussions on each design are given in Appendices C-1 through C-5.

4.3 INDOOR UNIT

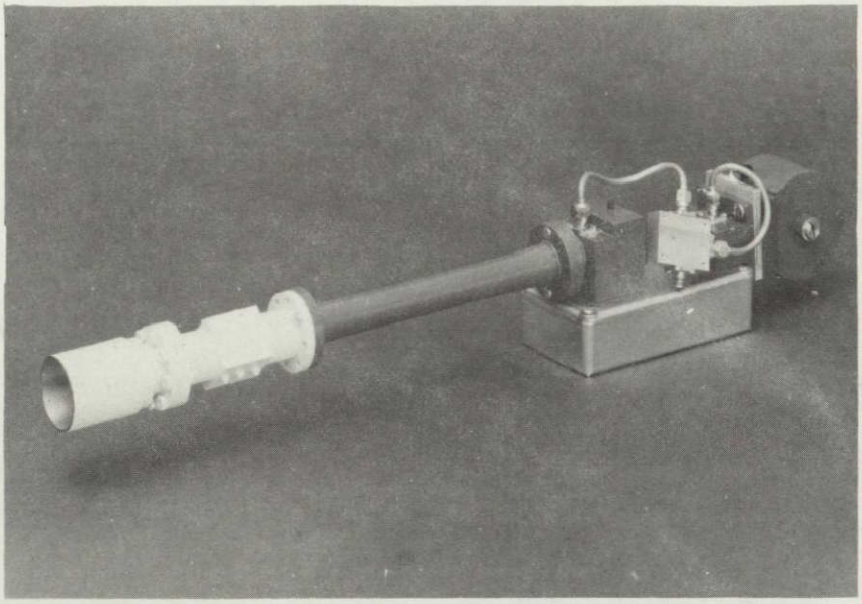
The indoor unit which is common to the 2.6 GHz and 12 GHz receivers consists of a second IF amplifier circuit, a limiter, a FM discriminator, a remodulator, a voltage regulator, and a power supply. The output signal for the direct reception is a VSB·AM-FM signal which is received by the conventional TV set. A VSB filter is not included in this simple receiver design for the home reception. The output signal for the community reception is a baseband video and a FM audio at 4.5 MHz which should be remodulated to a desired channel by cable TV operators. The remodulator circuit used is a simplified low-cost circuit, which is adequate for home reception. Detailed discussion on the indoor unit design is given in Appendix C-6.

4.4 THE ANTENNA

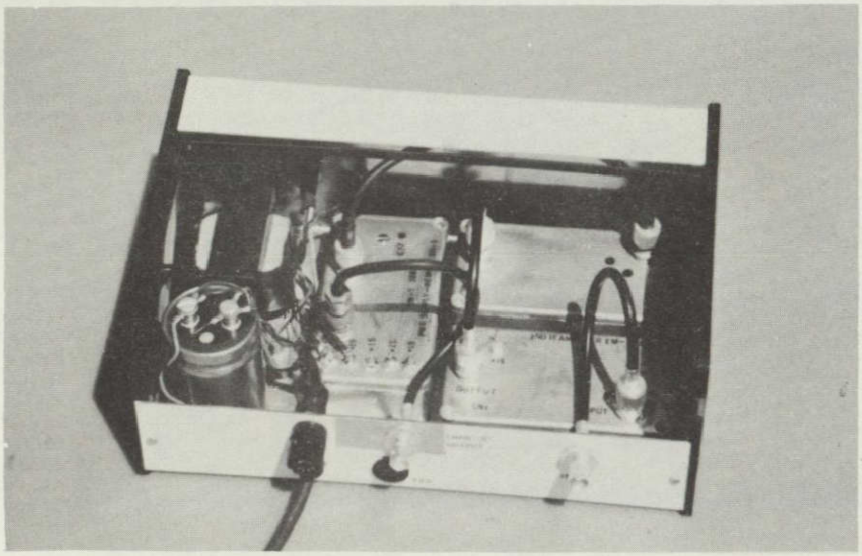
A 5 ft. to 10 ft. parabolic antenna is installed outside with the described outdoor unit at its focal point. Side-lobe suppression of the antenna is as important as high efficiency in order to avoid the interference from other satellites. Low-side lobe techniques have been investigated and are discussed in[21].

4.5 OVERALL PERFORMANCES OF WORKING MODELS

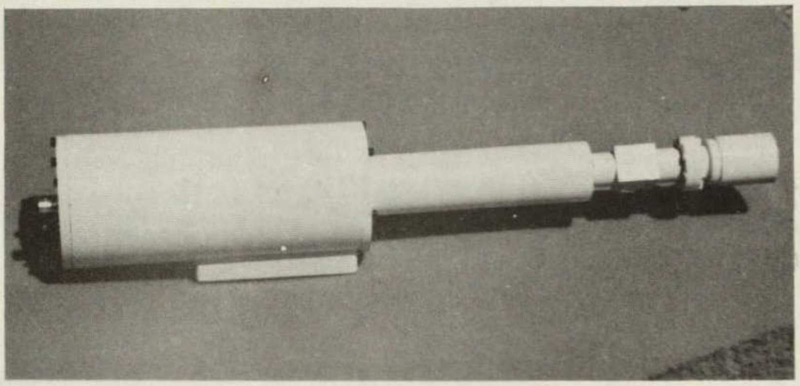
Two working models were built where the packagings were machined instead of diecast. Pictures 1 and 2 show the configuration of the outdoor unit and the indoor unit. Pictures 3 and 4 show the overviews after being packaged. Table 4-1 summarizes the overall performances.



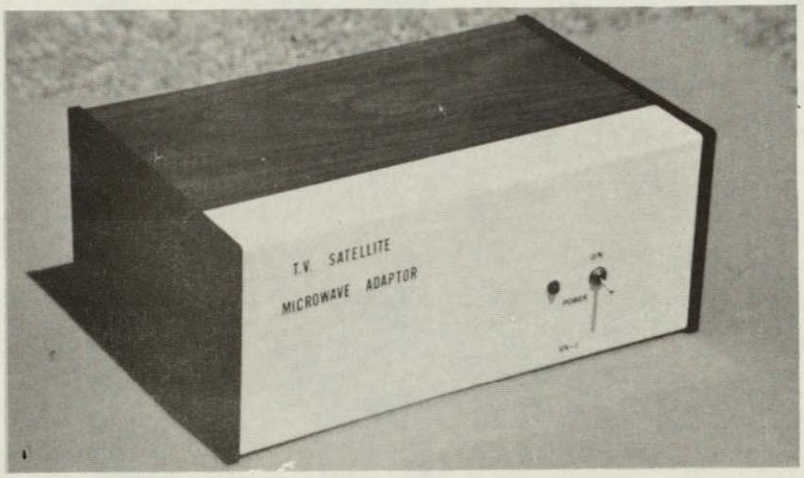
Picture 1. Working model - 12 GHz outdoor unit (from left to right: multi-mode horn, polarizer, high-pass filter, mixer, Gunn oscillator main cavity and invar stabilizing cavity; bottom: first IF amplifier packaging)



Picture 2. Working model - indoor unit (left: power supply, center: voltage regulator and remodulator, right top: limiter and discriminator and right bottom: second IF amplifier)



Picture 3. Overview of the 12 GHz outdoor unit



Picture 4. Overview of the indoor unit

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Table 4-1

Overall Performance of Working Models

	Specification	Measured Value
Outdoor Unit		
Noise figure		
System noise figure	10 dB	8.0 dB
Feed transmission loss		.5 dB
Mixer diode conversion loss		5.5 dB
1st IF amplifier noise figure		2.0 dB
Input voltage standing wave ratio	1.4	1.25
Image rejection ratio	15 dB	15.0 dB
Polarizer cross coupling		30.0 dB
axial ratio		1.0 dB
Frequency stability -30 +70°C 0~100% humidity 500~800 mmHG	$\pm$ 2 MHz (sealed)	$\pm$ 2.5 MHz (unsealed)
Gain		
Total Gain	30 dB	34.0 dB
RF loss		6.0 dB
1st IF amplifier gain	37 dB	40.0 dB
Mixer dynamic range		
IF bandwidth		40.0 MHz
Indoor Unit		
Noise figure		3.0 dB
2nd IF amplifier gain		40.0 dB
Limiter dynamic range		5.0 dB
Discriminator differential gain		$\pm$ 15.0%
differential phase		20.0 msec
Output voltage:		
Baseband		1.2 mV_{rms}
Remodulator output voltage		10.0 mV_{rms}
Output frequency:		
Baseband		0 ~ 4.5 MHz for video 4.5 MHz for audio
Remodulator output frequency		76~82 MHz

CHAPTER V

SYSTEM COST/PERFORMANCE ANALYSIS

5.1 GROUND-TERMINAL OPTIMIZATION

The ground-terminal consists of an antenna and a converter. The converter, here, consists of an outdoor unit and an indoor unit and does not include a television set. In Chapter III the 12 GHz converter design is optimized and the cost for a 10 dB noise figure converter analyzed. The cost of preamplifiers (TDA, uncooled parametric amplifiers and cooled parametric amplifiers) were studied by General Electric Company[22], Philco-Ford[7], and others[23]. These cost figures were used to get the converter cost vs converter noise figure for various manufacturing quantities per year. The same approach was applied to the 2.6 GHz converter design and to its cost estimation. The results are summarized in Fig. 5-1 where the selling price of a converter unit vs receiver noise figure is shown with production quantity (units per year) as a parameter.

The price of the antenna including installation cost was also studied[20][21] and summarized in Fig. 5-2. (55% efficiency is assumed.)

The figure of merit of the ground-terminal G_r/T_s is expressed by

$$G_r/T_s = G_r/(T_a + T_r) \qquad (5\text{-}1)$$

where G_r = receiving antenna gain

T_s = system noise temperature

T_a = antenna noise temperature (50°K for 2.6 GHz and 100° for 12 GHz)

T_r = receiver noise temperature

as was stated in Chapter II.

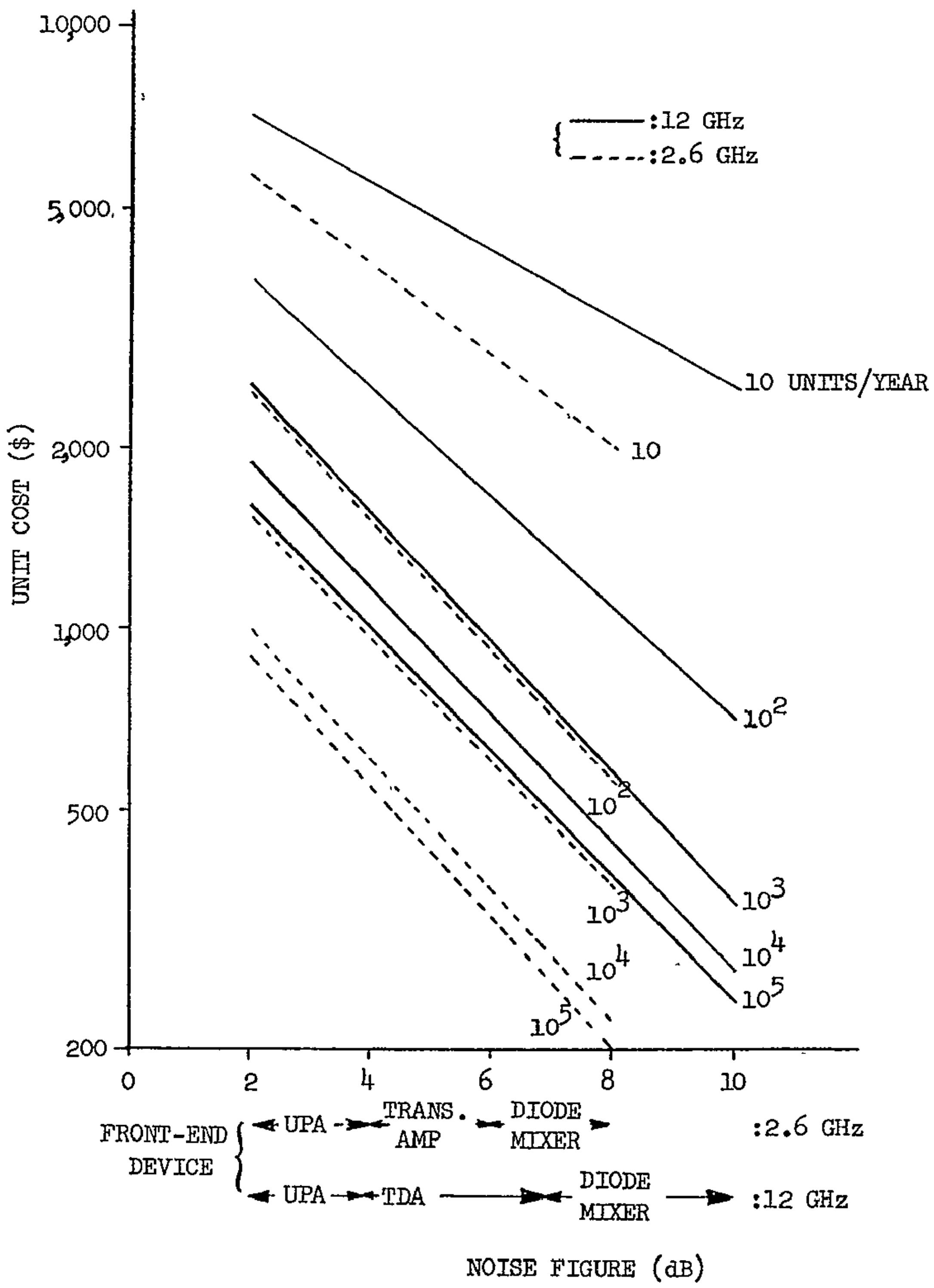


Figure 5-1 Receiver Selling Price (factory cost by 3) vs Noise Figure at Various Production Quantities for 2.6 and 12 GHz Systems

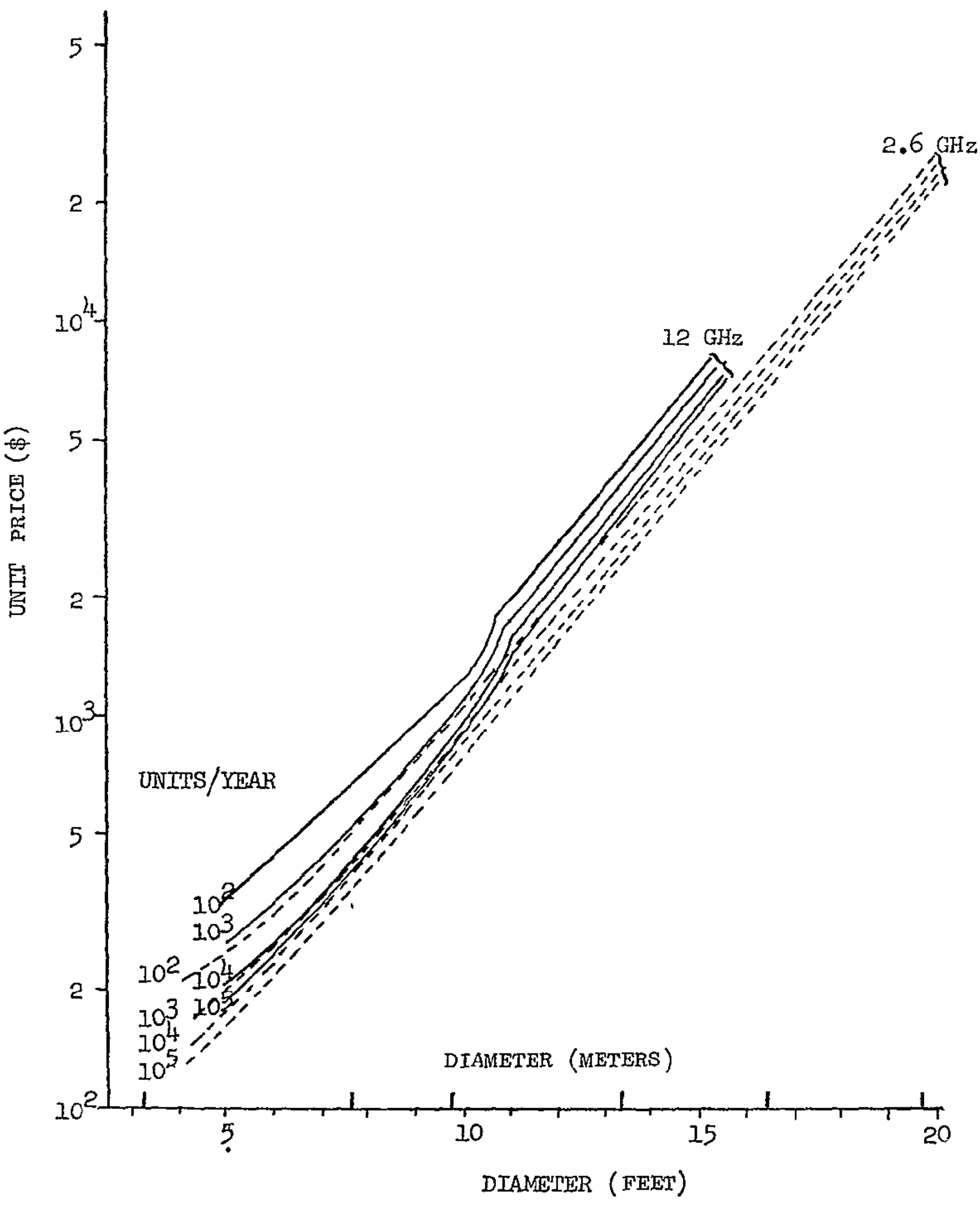


Figure 5-2 Installed Antenna Price vs Diameter for 2.6 and 12 GHz Systems

Optimization in G_r/T_s is done to minimize the cost of the ground-terminal by using Figs. 5-1 and 5-2 graphically and is summarized in Figs. 5-3 and 5-4 for both the 2.6 and 12 GHz systems. In Fig. 5-3, the optimum combination of the receiver noise figure NF_r and the antenna diameter for a given ground-terminal figure of merit G_r/T_s is shown for various production quantities per year. Figure 5-4 shows the resulting price of the optimized ground-terminal vs G_r/T_s.

From Fig. 5-3, it is advantageous to have a lower noise receiver and a smaller diameter antenna to get a larger G_r/T_s. This trend is more remarkable for larger production quantities. After reaching the minimum receiver noise figure obtainable (2.0 dB is assumed), the only way to increase G_r/T_s is then to increase the antenna diameter which raises the ground-terminal cost very rapidly as can be seen in Fig. 5-4. This comes from the fact that a low-cost low-noise microwave receiver is mass-producible by using the current microwave circuit technology, while the antenna cost increases rapidly with increasing size because of the high cost of labor for installation. In addition, the price of an antenna larger than 10 ft. in diameter for the 12 GHz system becomes very expensive due to the necessary tracking mechanism.

5.2 RAIN ATTENUATION AND SATELLITE COVERAGE

One of the biggest differences between the 2.6 GHz signal and the 12 GHz signal is rain attenuation. The 2.6 GHz signal attenuation is negligible, while the 12 GHz signal attenuation is considerable with high precipitation. Figure 5-5 illustrates rain attenuation in the U.S.A. studied by COMSAT[6] for the case where the indicated attenuation

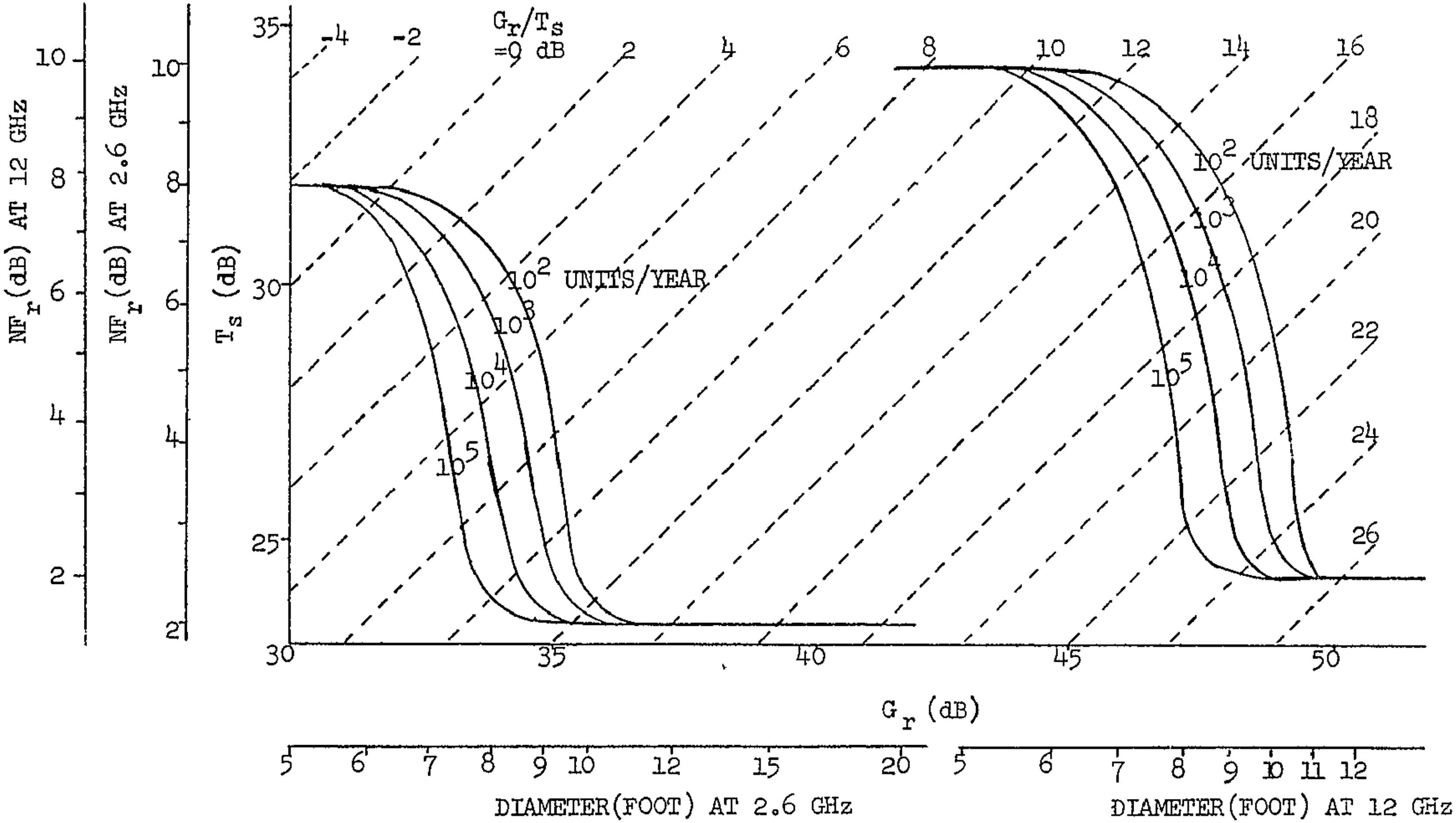


Figure 5-3 Optimum Ground-Terminal (combination of antenna and receiver) for the 2.6 and the 12 GHz Systems at Various Production Quantities Per Year and G_r/T_s (55% antenna efficiency and 50°K (100°K) of antenna temperature for the 2.61 (12) GHz are assumed)

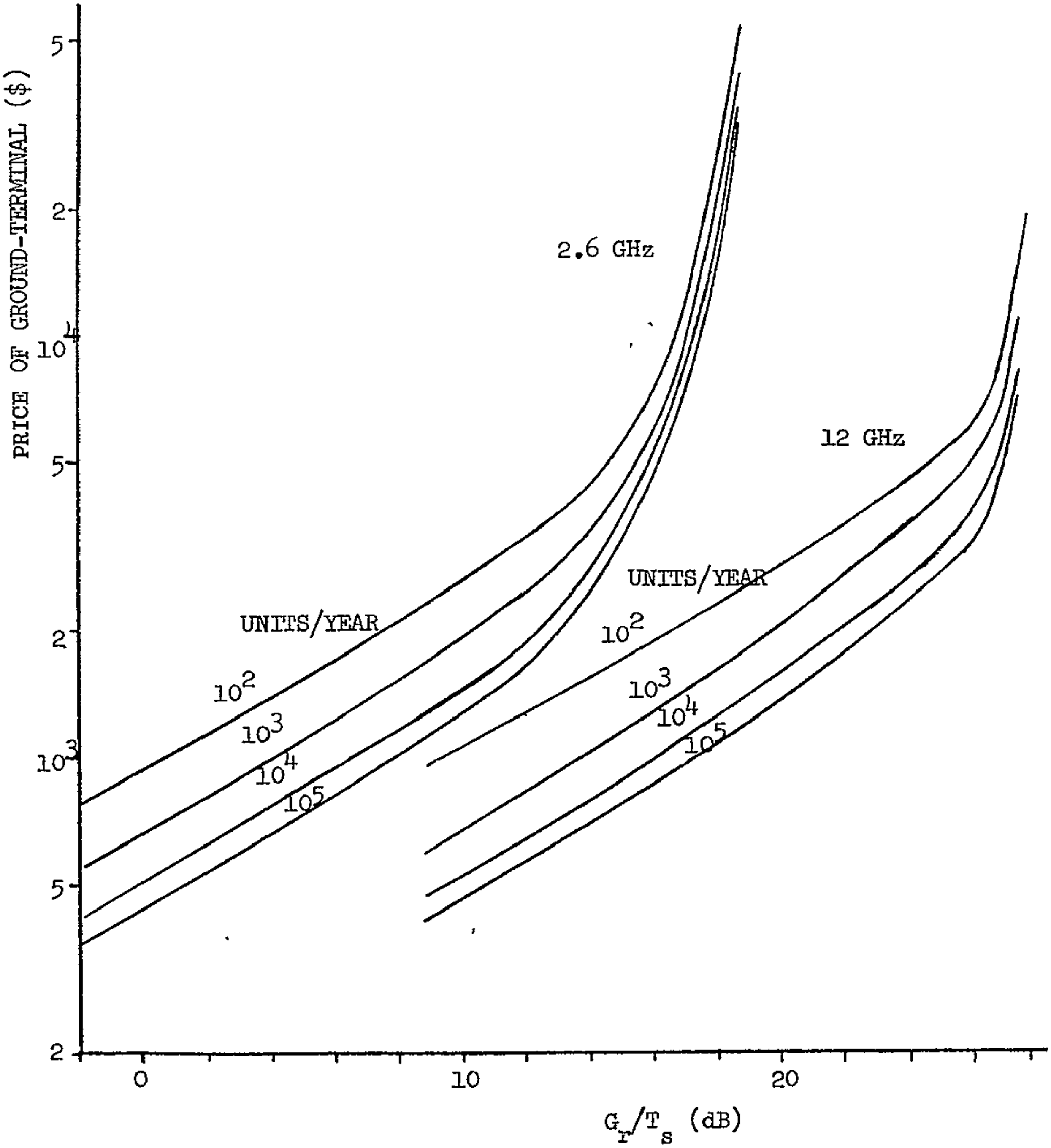
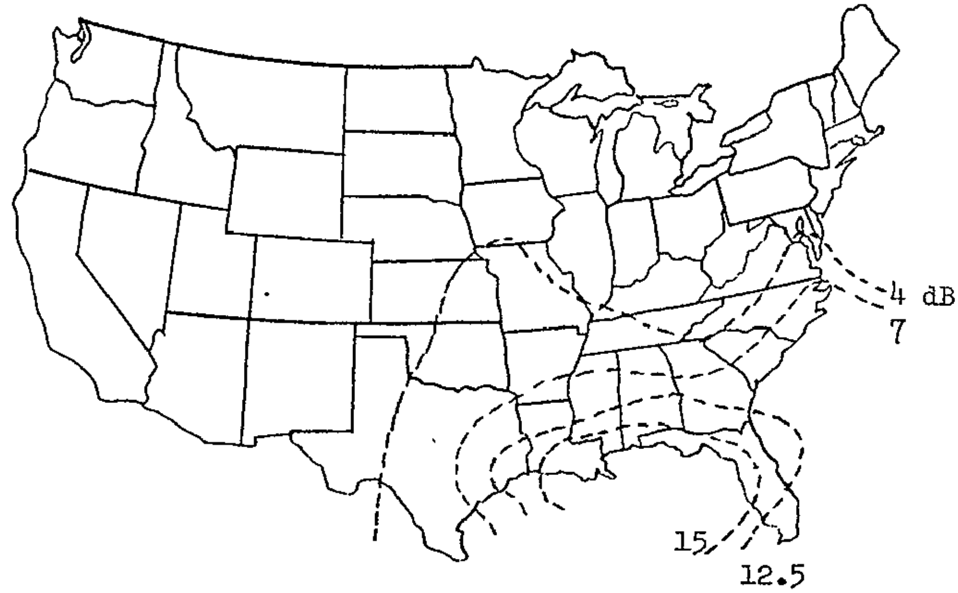
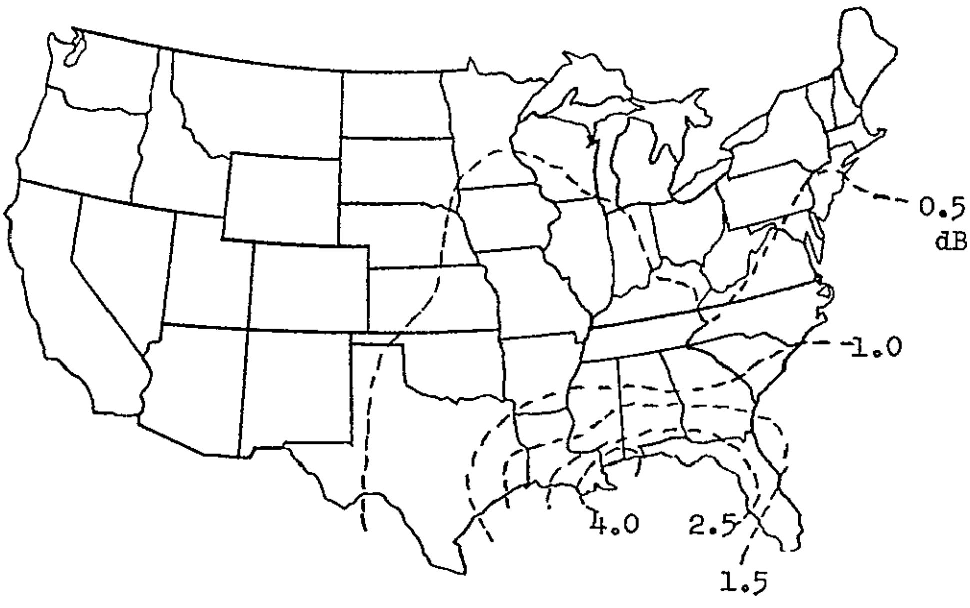


Figure 5-4 Optimized Ground-Terminal Selling Price vs G_r/T_s at Various Production Quantities for the 2.6 and the 12 GHz Systems



(a) 0.1 % OF OUTAGE



(b) 0.5 % OF OUTAGE

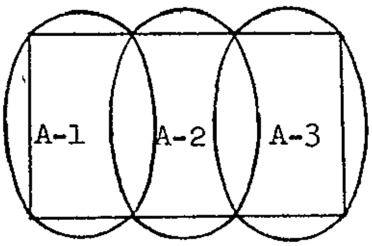
Figure 5-5 Rain Margin Map for the 12 GHz Signal for 0.1 and 0.5% of Outage (studied by COMSAT[6])

does not exceed 0.1% and 0.5% of the time per year, respectively. New Orleans area has the most attenuation where 0.05% value is 20 dB. But even there the 1% value is only 2 dB. In other words, if a guarantee of 99.95% of service time (0.05% of service outage) is made, up to 20 dB rain margin is needed, while for 99% of service time (1.0% of service outage) no more than 2 dB rain margin is needed in that area. Therefore, careful consideration of the satellite coverage, satellite transmitter power, ground-terminal G_r/T_s and service time to optimize the cost/ performance of the 12 GHz system is needed.

Figure 5-6 shows two (and there are many others) alternative satellite coverages for the U.S.A. Basically, the country is divided into three regions to tolerate the time differences. The alternative [A] has three identical $2.7^{\circ} \times 3.5^{\circ}$ beams which is to be used for the 2.6 GHz system and the inexpensive and low quality 12 GHz systems. The alternative [B] has one $2.7^{\circ} \times 3.5^{\circ}$ beam and four $2.7^{\circ} \times 1.75^{\circ}$ beams which are to be used for high quality 12 GHz systems.

5.3 COST OF SATELLITE SEGMENT

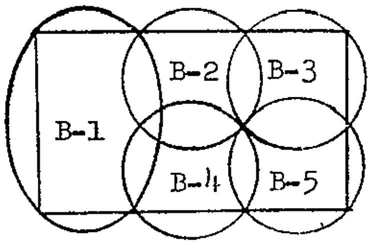
In the past decade, the technology has been considerably improved to increase the power available in satellites from a few watts in the early commercial satellite to 672 watts which is the goal of the MCI Lockheed spacecraft[24]. Research and development is taking place to increase this figure to the kilowatt (kw) region[25][28]. For the present analysis the available rf power is restricted to 1 kw.



3 dB BEAMWIDTH=$2.7^\circ \times 3.5^\circ$

G_{sat}=34.6 dB

(a) ALTERNATIVE [A]



3 dB BEAMWIDTH= $2.7^\circ \times 3.5^\circ$ FOR B-1
$2.7^\circ \times 1.75^\circ$ FOR B-2,3,4 & 5

G_{sat}= 34.6 dB FOR B-1
37.6 dB FOR B-2,3,4, & 5

(b) ALTERNATIVE [B]

Figure 5-6 Alternative Satellite Coverages

The rf output power P_{sat} is related to EIRP in the following manner:

$$P_{sat} = EIRP - G_{sat} + L_{sat}$$

where EIRP = effective isotropic radiated power

G_{sat} = gain of the satellite antenna (55% of efficiency is assumed).

L_{sat} = loss between the power amplifier and the feed, 1.6 dB

The gain of the antenna G_{sat} is related to the 3 dB beamwidths θ_1 and θ_2 (in degrees) as

$$G_{sat} = 10 \times \log \frac{27{,}000}{\theta_1 \times \theta_2} \tag{5-3}$$

The cost of one satellite segment including the up-link facility, space vehicle, launching and annual operating and maintenance can be measured by its rf output power. The recent papers show that the cost of one watt ranges from \$35,000 to \$60,000 per year and is fairly constant up to 100 kw output power[10][24]. In the numerical cost/performance analysis to follow, it is assumed that the cost of the satellite segment is determined by the total rf output power and the cost of one watt is \$50,000 per year. It is also assumed that the manufacturing capability of ground-terminals is not less than 10,000 units per year. The analysis will be carried out on the basis of the annual system cost.

5.4 SYSTEM COST/PERFORMANCE OF THE 2.6 GHz SYSTEM

The coverage alternative [A] is to be used for the 2.6 GHz system. From the calculations in Chapter II, the required EIRP is

expressed by

$$\mathrm{EIRP} = \mathrm{CNR} - (G_r/T_s) + 45.9 \quad (\mathrm{dB}) \qquad (5\text{-}4)$$

for the 2.6 GHz FM system with 25 MHz rf signal bandwidth. The gain of the satellite antenna G_{sat} is 34.6 dB from Eq. (5-3) for $\theta_1 = \theta_2 = 3^\circ$. Hence, from Eq. (5-2) the rf output power P_{sat} for each region is

$$P_{sat} = \mathrm{CNR} - (G_r/T_s) + 11.3 \quad (\mathrm{dB}) \qquad (5\text{-}5)$$

The total annual system cost COST_{system} is then given by

$$\mathrm{COST}_{system} = 150{,}000 \times P_{sat\ watts} + \mathrm{COST}_{gt} \times Q \quad (\$) \qquad (5\text{-}6)$$

where COST_{gt} is the annual unit cost of a ground-terminal and Q is the total quantity of the gound terminals. Here it is assumed that the ground-terminals are amortized over 10 years at an interest rate of 10% and thus, the annual payments on capital investments amount to 16.2% of the first cost. In addition, it is also assumed[10] that the estimated annual cost of operations and maintenance of ground-terminals is 25% of the first cost. Hence, the annual cost of a ground-terminal is 41.2% = 25 + 16.2 of the first cost given in Fig. 5-4. The minimized system cost COST_{system} and the corresponding P_{sat} and G_r/T_s for various quantities Q are shown in Fig. 5-7 for the home reception system with a SNR of 45 dB and 42 dB. From the figure, higher satellite transmitting power and lower ground-terminal G_r/T_s are advantageous, with the increase in the number of ground-terminals.

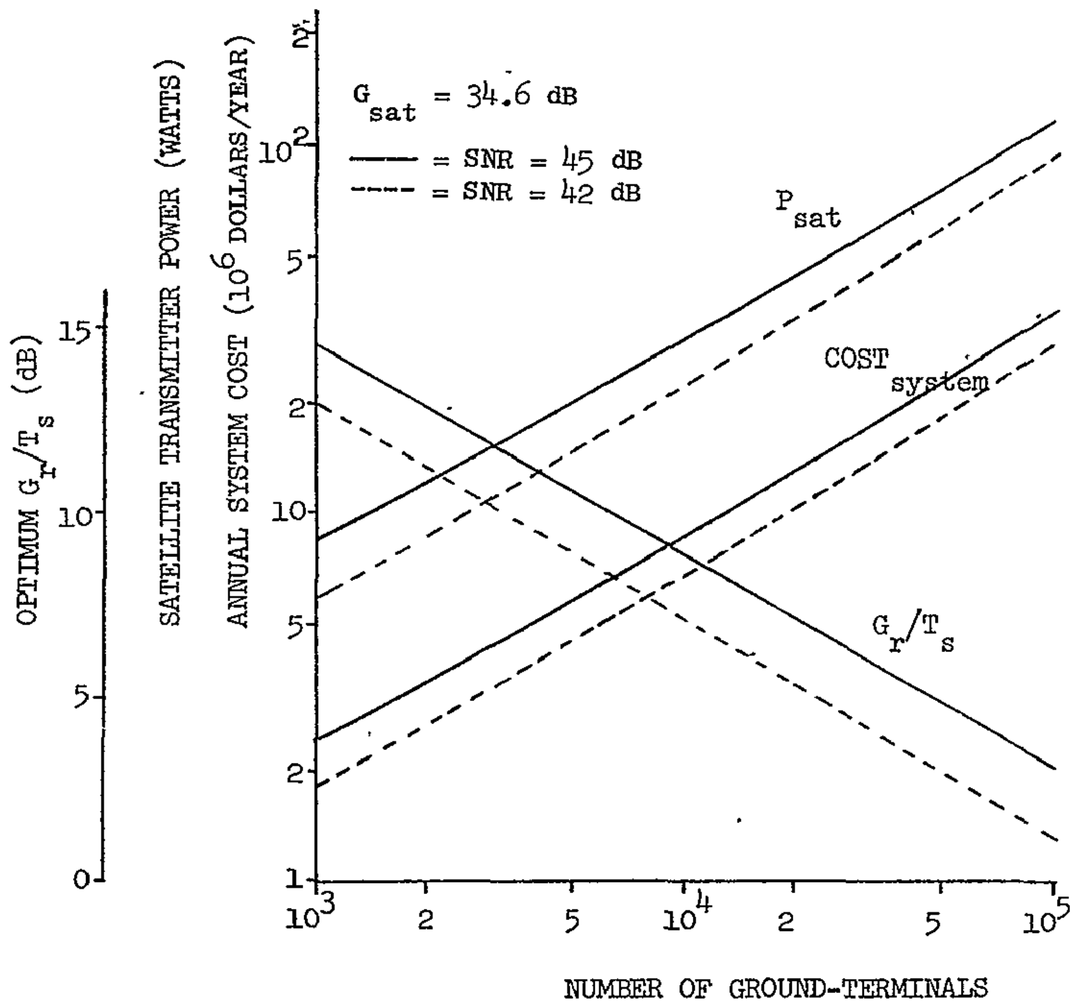


Figure 5-7 Annual System Cost, Satellite Transmitter Power and Ground-Terminal G_r/T_s at the Optimum for Various Numbers of Ground-Terminals for the 2.6 GHz System

5.5 SYSTEM COST/PERFORMANCE OF THE 12 GHz SYSTEM

The attenuation of the 12 GHz signal depends on the precipitation and hence on the location. Therefore, the need for more satellite transmitter power P_{sat} and/or more ground-terminal figure of merit G_r/T_s, in other words, more rain margin, for areas where more precipitation is encountered. After all, the 12 GHz system has some percentage of service outage time inevitably. The outage is defined here as the percentage of time when the rain absorption exceeds the rain margin. The following assumes that all other system margins are lost except for the rain margin.

The object of this section is to investigate the relationship between the rain margin, the service time (100% - outage), SNR at the ground-terminal, the ground-terminal G_r/T_s, the satellite transmitter power P_{sat} and the total annual system cost $COST_{system}$ for a variety of numbers (Q) of the ground-terminal.

Figures 5-5 and 5-6 show that there is quite a lot of signal attenuation for a certain small percentage of time in the south eastern area of the U.S.A. The relationship between the encountered rain attenuation and the percentage of locations for the various percentages of outage is shown in Fig. 5-8. In an extreme case, if 99.99% of service time (0.01% of outage), is needed more than 20 dB of rain margin is required for 8% of the locations. Consequently, the system cost is strikingly affected by how much service time one wants.

In this analysis, the interest is particularly in the low-cost direct broadcasting system so that more than 99.9% of service time (less than 0.1% of outage) is not of interest. For the higher quality

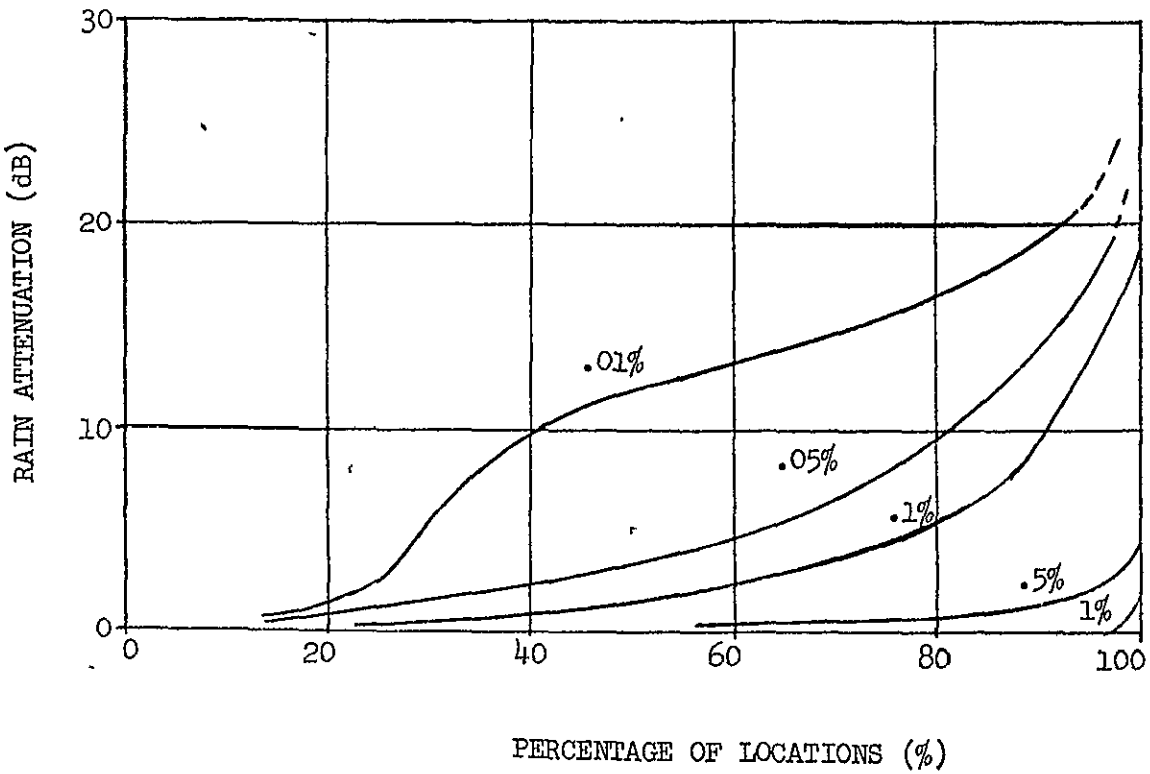


Figure 5-8 Rain Attenuation vs. Percentage of Locations in the U.S.A. for the 12 GHz Signals Broadcast from Satellites (Studied by COMSAT[6]. Parameter is percentage of outage)

service which may be required for the public broadcasting system, CATV redistribution system and so on, the space diversity technique used to send high quality signals received at a non-raining town to a raining town using a terrestrial microwave link like the cable relay system (CARS) band will be a powerful method to increase the system quality.

In the following sections, rain margin, satellite coverage, ground-terminal G_r/T_s, satellite transmitter power, and the annual system cost for the three cases of [I] 1.0%, [II] 0.5%, and [III] 0.1% of outage are discussed.

CASE I - 1.0% outage

2 dB rain margin is needed for 5% of the locations in the New Orleans area. This margin can practically be achieved by installing a slightly bigger antenna for this area. The coverage alternative [A] is used.

According to the calculations in Chapter II, the required EIRP for the 12 GHz system is expressed by

$$\mathrm{EIRP} = \mathrm{CNR} - (G_r/T_s) + 59.2 \quad (\mathrm{dB}) \qquad (5\text{-}7)$$

where the signal bandwidth of 25 MHz is assumed. The gain of the satellite antenna G_{sat} is 34.6 dB so that from Eq. (5-2), the rf satellite power P_{sat} is

$$P_{sat} = \mathrm{CNR} - (G_r/T_s) + 24.6 \quad (\mathrm{dB}) \qquad (5\text{-}8)$$

The equivalent annual cost of the ground-terminals $COST_{gt}$ for CASE I is given by

$$COST_{gt} = 0.95 \times COST_{gto} + 0.05 \times COST_{gto + 2\ dB} \qquad (5\text{-}9)$$

where $COST_{gto}$ is the annual cost of the ground-terminal which is used for 95% of the location and $COST_{gto + 2\ dB}$ is the annual cost of a ground-terminal which has 2 dB higher G_r/T_s and covers 5% of the location. The total annual system cost $COST_{system}$ is given by Eq. (5-6). Here the assumption is made that the ground-terminals are uniformly distributed throughout the country.

The total annual system cost $COST_{system}$, the satellite transmitter power P_{sat} and the ground-terminal G_r/T_s at the optimized design are shown in Table 5-1.

CASE II - 0.5% of outage

The coverage alternative [A] is used with different G_r/T_s depending on the location to achieve at most 0.5% of outage all over the country as was done for CASE I. From Fig. 5-9, G_r/T_s of 2 dB larger is used for 20% of the location and G_r/T_s of 4 dB larger is used for 10% of the location. Then the equivalent annual cost of the ground-terminal $COST_{gt}$ for CASE II is

$$COST_{gt} = 0.7 \times COST_{gto} + 0.2 \times COST_{gto + 2\ dB} + 0.1 \times COST_{gto + 4\ dB} \qquad (5\text{-}10)$$

Table 5-1

The Minimum Annual System Cost $COST_{system}$, Satellite Transmitter Power P_{sat} and Ground-Terminal G_r/T_s for the Three Cases, CASE I, CASE II and CASE III (12 GHz)

CASE	Outage (%)	Coverage	Rain Margin (dB)	% of Location	Number of Ground-Terminals											
					10^3				10^4				10^5			
					P_{sat} (w)	G_r/T_s (dB)	$COST_{system}$ ($million)		P_{sat} (w)	G_r/T_s (dB)	$COST_{system}$ ($million		P_{sat}	G_r/T_s	$COST_{system}$ ($million)	
I	1.0	[A]	0	95	11.5	26.0	3.4		42	20.5	11.6		164	14.5	47	
			2	5		28.0				22.5				16.5		
II	0.5	[A]	0	70	20.5	23.5	4.5		46	19.5	14.5		184	14.0	57	
			2	20		25.5				21.5				16.5		
			4	10		27.5				23.5				18.5		
III	0.1	[B]	3	33.3 (B1)	22.5	26.5	2.0	26	50	22.5	5.2	45.6	200	16.5	21	109
			7	33.3 (B 2&3)	18.0 each	28.0	4.0		40 each	24.5	8.4		144 each	19.0	28	
			16	33.3 (B 4&%)	92.0 each	30.0	20		144 each	28.0	32		288 each	25.0	60	

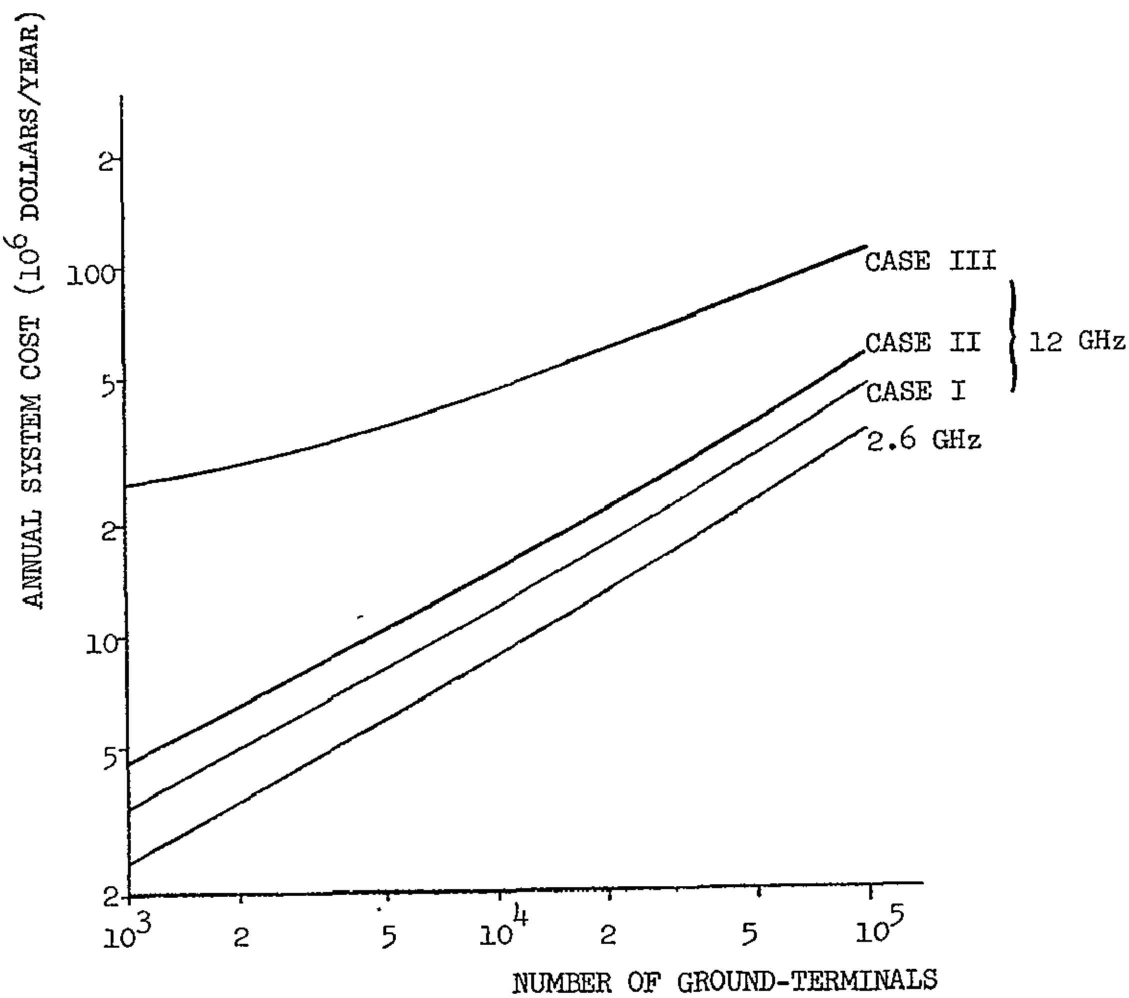


Figure 5-9 Annual System Cost vs Number of Optimum Ground-Terminals for the 2.6 GHz and the 12 GHz Systems

Now the minimized $COST_{system}$, P_{sat} and G_r/T_s are calculated from Eqs. (5-6), (5-7), (5-8) and (5-10) and are summarized in Table 5-1.

CASE III - 0.1% of outage

To achieve 0.1% of outage a 16 dB rain margin is needed. This is indicated in Figs. 5-5 and 5-8. The amount of required rain margin varies depending on the location that the coverage alternative [B] must use. Then, the needed rain margins are 3 dB in the region B-1, 7 dB in the regions B-2 and B-3 and 16 dB in the regions B-4 and B-5.

Using similar procedures, the optimization of the annual system cost $COST_{system}$, the satellite transmitter power P_{sat} and the ground-terminal G_r/T_s was done for the five regions, the results of which are summarized in Table 5-1.

The annual system cost of the 2.6 GHz system and the three cases of the 12 GHz system is depicted in Fig. 5-9 for comparison.

5.6 THE ACTUAL SIGNAL QUALITY EVALUATION FOR THE 12 GHz SYSTEM

So far, the "worst case" signal quality where all the possible signal degradations were taken into account has been discussed. In addition to the rain margin discussed in the preceeding section, the following margins are included: antenna pointing error $L_{pointing}$ (1.0 dB), atmospheric absorption and long-term system degradation, L_{atom}(0.4 dB), polarization loss, L_{pol} (0.1 dB), design margin for unknowns, M(1.0 dB), and off-beam center margin, L_{beam} (3.0 dB). Among these five margins, the first four terms (2.5 dB to sum up) are margins for probabilistic phenomena and the last term (3 dB) is the margin to get the specified

signal quality at the beam-edge. In other words, the first four depend on the time, while the last depends on the satellite coverage and the location of the ground-terminal. It is reasonable to assume that the first four are caused by random phenomena and hence this 2.5 dB margin is distributed according to the Poisson distribution. Therefore, the average SNR will be about 1.7 dB better than the worst case SNR dealt with thus far.

Outage is defined as the percentage of time when the received worst case SNR will be below 45 dB or the corresponding CNR will be below 11 dB. It is important to note that even below the FM threshold (11 dB), SNR does not abruptly go down but some picture quality is still maintained. Figure 5-10 shows the measured relation between SNR and CNR.

Figure 5-11 shows the whole figure of the outage vs SNR for the three Cases I, II, and III at New Orleans, the area where the most rain attenuation is encountered in the U.S.A. and the 3 dB off-beam margin has to be counted, in other words, where the poorest picture quality in the country will be received. The figure shows the range of the picture quality for the three cases. The best quality occurs where no probabilistic attenuation except the rain attenuation is encountered (subscripted by B) and the worst quality occurs where all the probabilistic attenuation except the rain is encountered (subscription by W).

5.7 DISCUSSIONS ON COST/PERFORMANCE

Studies have been done on the minimized annual system cost, the satellite transmitter power and the ground-terminal G_r/T_s for the 2.6 GHz system, and the 12 GHz system. Three cases were studied for the

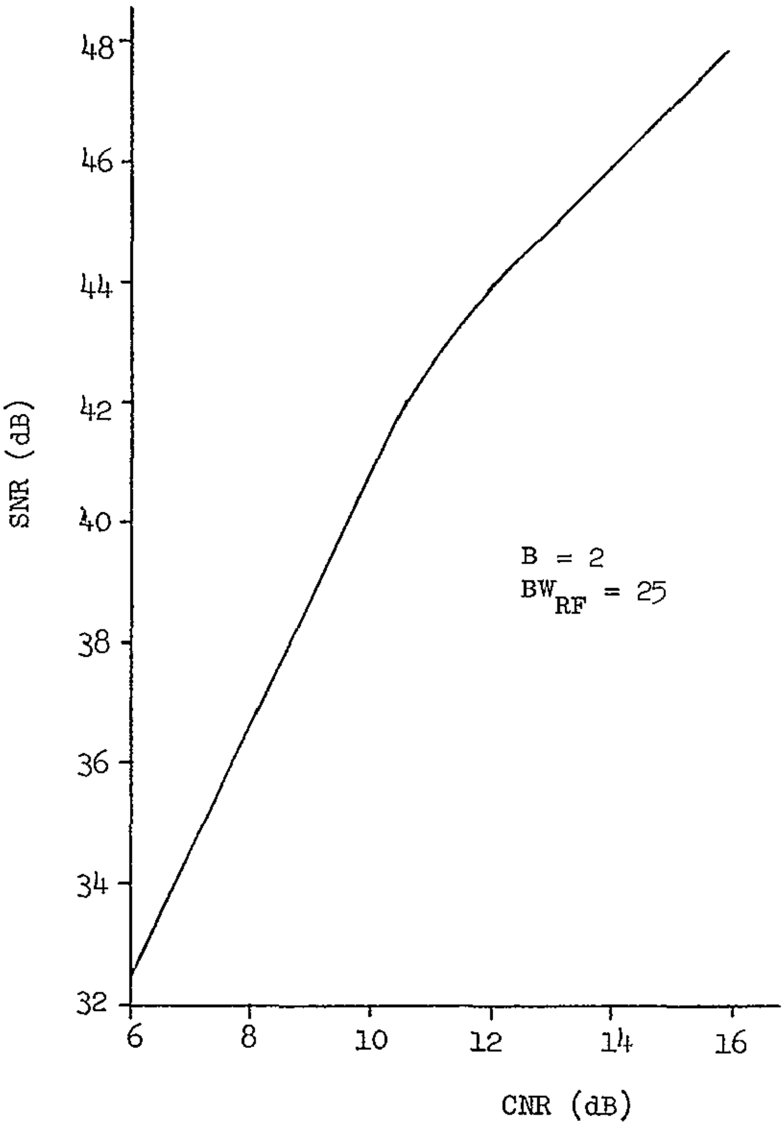
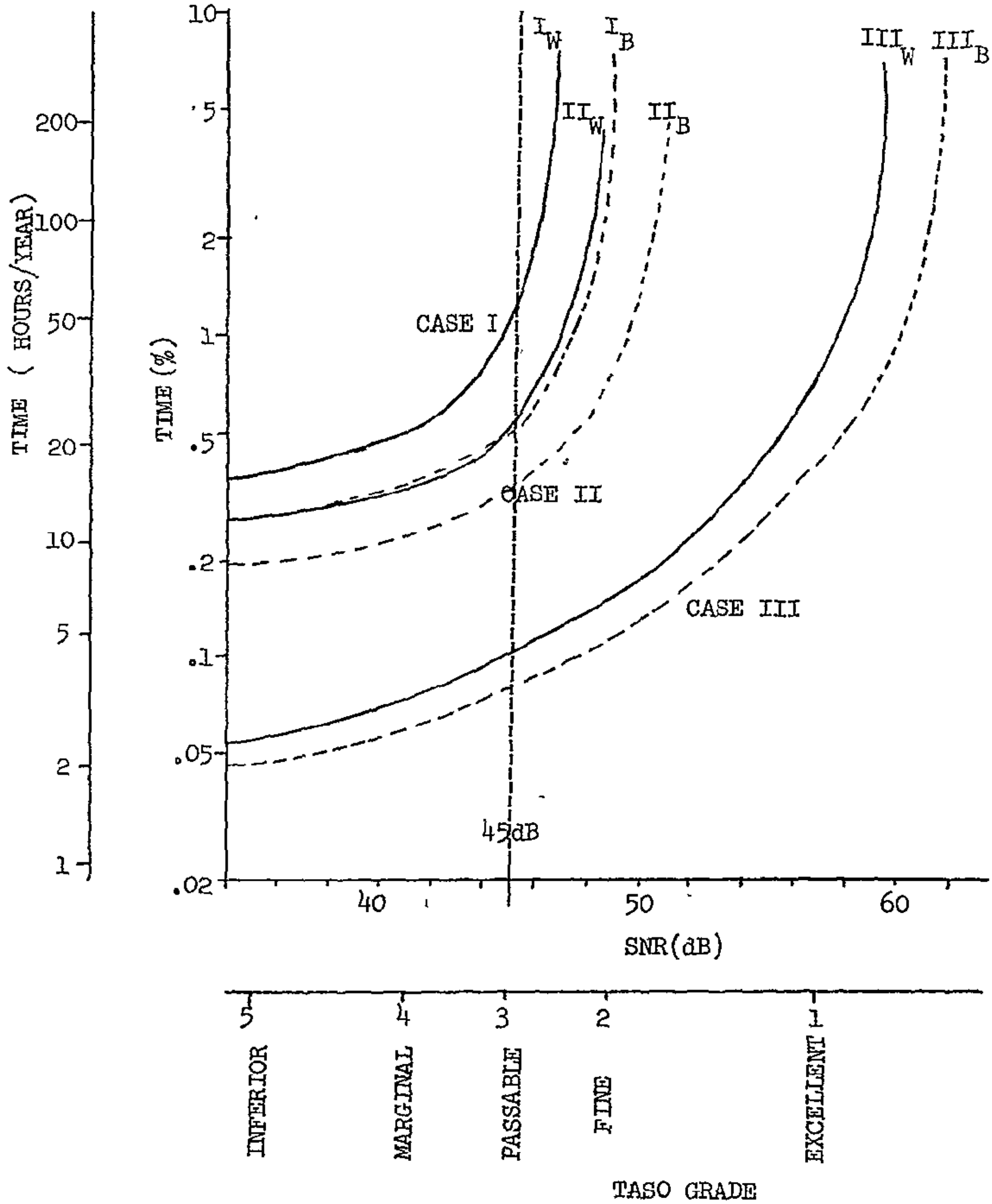


Figure 5-10 Measured SNR vs CNR



The solid lines show the best cases where probabilistic degradation except rain attenuation is not encountered (noted by B). The dotted lines show the worst case, on the other hand (noted by W). For outage; 12 hours a day of service is assumed. (12 GHz).

Figure 5-11 Time (% and hours/year) vs Picture Quality for CASE I, II, and III and for the 12 GHz System at New Orleans which has the most Rain Attenuation

12 GHz system; CASE I (1% of outage), CASE II (0.5% of outage) and CASE III (0.1% of outage). CASE I is the lowest cost system which has no rain margin for 95% of the locations and a 2 dB margin for the remaining 5%. On the average of time and at the worst location in the country (the New Orleans area), one gets 45 dB of SNR (TASO grade 3 --- "passable" as shown in Fig. 5-11) for 99.3% of the time (an outage of 30 hours a year if 12 hours a day of service is assumed) and 35 dB of SNR (TASO grade 5 --- "inferior") for 99.7% of the time (an outage of 15 hours a year). By using CASE II, the system cost is raised 20 to 30% but the outage from 45 dB SNR becomes 15 hours a year and the outage from 35 dB SNR becomes 10 hours a year. But if 0.1% of outage is needed (CASE III) which gives 3.5 hours a year of 45 dB outage and 2 hours a year of 35 dB outage, the system cost is 10 times for a quantity Q of 10^3 and 3 times for a quantity Q of 10^5 as much as CASE I.

In Chapter II, a discussion of a desirable SNR for the broadcasting satellite was given and in it a conclusion was reached that stated 45 dB (TASO grade 1.5) would be necessary for the community reception system. The latter needs 9 dB more G_r/T_s than the former. The optimization of the ground-terminal was discussed in Section 5-1. Now, the next question is "how small a percentage of outage should be used for the 12 GHz system?".

Think about an application to education for example, which is one of the most important fields of application of broadcasting satellites. Ideally, any class should not be cancelled, but it usually happens that several classes in a year are cancelled due to illness of the instructor, bad weather, crowded traffic, or some other reason. Talking about CASE II, 10 hours a year of complete dismissal and an additional 5 hours a

a year of unsatisfactory picture quality could be tolerable. CASE I has about twice as much outage which is believed to be of a tolerable limit for educational purposes even though it has 20 to 30% less cost. CASE III is too costly in realization for this purpose. If the programs are distributed through cable TV networks by community reception, the outage is remarkably reduced by the use of space diversity which connects the main ground-terminals by terrestrial microwave links using CARS band or other frequency bands. The use of this diversity is much less costly than having more rain margin as for CASE III.

Consequently, it is concluded that CASE II will give the best cost/performance for the low-cost direct reception system using the 12 GHz band at least for educational purposes. The cost of the 12 GHz system using CASE II is about 60-80% more than the 2.6 GHz system.

CHAPTER VI

6.1 CONCLUSION

A direct reception system from broadcasting satellites has been studied, technically and economically, with emphasis on utilization of the 12 GHz band. From a technical standpoint both the 2.6 GHz system and the 12 GHz system are already possible by fully utilizing current highly developed technologies. The 2.6 GHz system is going to be tested using the ATS-F satellite by the Federation of Rocky Mountain States in 1974. This study consists of (1) technologies for 12 GHz low-cost ground-terminals, (2) communication systems suitable for direct reception and (3) system cost/performance analysis. The results are reviewed and summarized below

(1) Technologies for the 12 GHz low-cost ground-terminals

1-1 The operation mechanism of the cavity stabilized negative-resistance diode oscillator was made clear and the effects of circuit parameters were analyzed. High Q of the stabilizing cavity and tight coupling between the main cavity and the stabilizing cavity is preferred.

1-2 The stabilizing cavity must be made by a low-thermal expansion material like invar and must be hermetically sealed. Frequency stability of $\pm$ 1.0 x 10^{-4} or better is available for $\pm$ 50°C of temperature range and 0~100% of humidity range. The effect of air pressure change is negligible.

1-3 Invar cavities can be either cast or machined out of a tube and a plate.

1-4 A feed consisting of a horn, polarizer and a high-pass filter needs little or no adjustment.

1-5 The rest of the circuitry should be made by integrated circuits in case of large quantities.

1-6 Outdoor units are to be die cast for low-cost in case of mass-production. An invar cavity with the inside surface silver-plated is attached to the cast aluminum packaging by screws.

1-7 The outdoor unit is made compact, protected against weather and installed at the focus of a parabolic reflector.

1-8 The tested results of the working models were satisfactory.

(2) Communication systems suitable for direct reception

2-1 FM is the best alternative modulation method for broadcasting satellites both from the limitation of satellite power and possible interferences to terrestrial communication systems.

2-2 AM-FM was analyzed. Although it needs a simpler receiver configuration, the required transmitter power from satellites is about 10 to 15 dB more than FM, which makes AM-FM practically impossible at least today and in the near future.

2-3 The home reception will need a SNR of 45 dB. On the other hand, the community reception will need 54 dB which is achieved by using 9 dB more ground-terminal G_r/T_s than home reception.

(3) System cost/performance analysis

3-1 The cost of a 12 GHz receiver is from \$350 to \$240 for 10 dB NF or from \$750 to \$500 for 7 dB NF corresponding to manufacturing quantities of 10^3 units per year to 10^5 units per year, the cost of which is about 60-90% higher than a 2.6 GHz receiver.

3-2 The ground-terminal was optimized for a given G_r/T_s.

3-3 System cost was minimized in terms of annual cost assuming \$50,000 per year, per watt[10],[24] for the satellite segment.

3-4 The design of the 2.6 GHz system is straight forward where there is negligible rain attenuation of signals. The required satellite power ranges from 8 to 100 watts and the ground-terminal G_r/T_s ranges from 15 to 3 dB corresponding to the number of ground-terminals ranging from 10^3 to 10^5 for 45 dB SNR. Corresponding annual system cost is \$3 million to \$25 million.

3-5 On the other hand, the 12 GHz system, which has a considerable amount of rain attenuation, needs careful evaluation of the required SNR, required service outage time and system cost. The proposed system (CASE II) for educational purposes has 15 hours a year of 45 dB outage or 10 hours a

year of 35 dB outage (if 12 hours a day of service time is assumed) The corresponding satellite power is from 20 to 184 watts and G_r/T_s is from 26 to 14 dB for the number of ground-terminals of 10^3 to 10^5. The satellite coverage is the same as the 2.6 GHz system which has three identical beams on the U S.A.. The corresponding annual cost is $4.5 million and $57 million, which is 60-80% higher than the 2.6 GHz system.

3-6 The community reception needs less outage which will be achieved by the use of terrestrial microwave links like CARS band.

3-7 It is the current tendency to go toward more channels for any application, so that modification to multi-channel TV transmission will be desired. Multi-channel TV transmission using multi-carrier FM raises the receiver cost only about 20% for a 12 channel transmission. The cost for the satellite segment for the multi-channel transmission increases with increasing the number of channels, as long as we measure the cost in terms of total rf output power from the satellites. In this case, the optimum system with minimum annual system cost has lower rf power from satellites and higher G_r/T_s for ground-terminals[10].

The discussion has been limited to a one-way broadcasting system In addition to a recent analysis of the economy of a two-way teleconferencing system[10], the analysis of the technical aspects and cost of an additional ground-terminal facility for the return links is recommended for future study to bring the 12 GHz broadcasting satellite into the daily convenience of nations.

The ATS-F satellite, which is to be launched in early 1974, is going to provide educational programs to the Rocky Mountain area to test a 2.6 GHz FM direct reception system for a period of one year. The program is also to be distributed through terrestrial cable TV links. No doubt the marriage of broadcasting satellites and cable TV links will give enormous communication means, providing nation-wide

programs for education, medical services, teleconferencing and so on, these benefits of which may be used to the social advantage of developed and developing nations.

In this research the technical and economical feasibility of the 12 GHz broadcasting satellite system, which would offer more channels and less interference problems than the 2.6 GHz system, has been studied. The cost/performance of both frequency bands has been analyzed and compared. The results of the study of these two frequency bands are so promising that it can be said that it is possible to see world-wide use of broadcasting satellite networks providing various kinds of information to all other nations in the coming one or two decades.

Appendix A

Frequency Stability Analysis of Cavity Stabilized Microwave Negative Resistance Diode Oscillator

Introduction

A-1 Requirements for Frequency Stability of Local Oscillator for Communication Use

It is true that negative resistance diodes such as Gunn diodes and IMPATT diodes are going to replace klystones and TWTs at microwave frequencies for low to medium power applications, but stabilization of the oscillating frequency of these solid state oscillators has remained as one of the serious problems to be overcome. The free-running frequency stability (stability is defined as frequency variation divided by the center frequency) of these diode oscillators is about $\pm$ 1.0 to 2.0 x 10^{-3} (corresponding to $\pm$ 10 to 20 MHz at 10 GHz) for the whole operational condition, (temperature, humidity, and pressure will be discussed in Section A-3) which is sufficient for some particular purposes, for instance, airborne radars and so on, but is quite inadequate for general communication purposes. Hence, some new developments in circuit technology are needed to get satisfactory stability performance.

Common techniques for the frequency stabilization of microwave oscillators are:

(1) injection locking by a stable frequency source

(2) automatic frequency control by using an AFC loop and a varactor diode

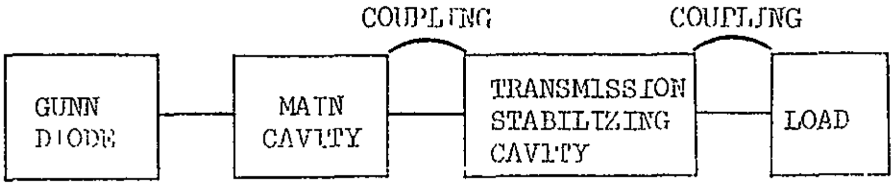
The first method uses a quartz oscillator and a step recovery diode circuit, which usually gives the best performance but is relatively costly. The second method is a good alternative for very low microwave frequencies,

but is not necessarily good for higher frequencies than 10 GHz because of the loss in the varactor diode. A third method is to use:

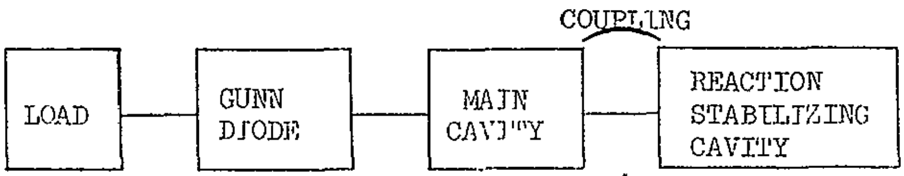
(3) a stabilizing cavity

The ideal cavity has a high Q and low thermal expansion. A transmission stabilizing cavity has been used to reduce oscillator noises (mainly FM noise) of klystrons. The block diagram is shown in Fig. A-1 (a). In 1968, J. R. Ashley et al. [25] used a high Q transmission stabilizing cavity for microwave negative resistance diode oscillators and obtained good noise reduction experimentally. It is easily understood that if a low thermal expansion material like invar is used, the long-term frequency fluctuation (frequency deviation caused by temperature variations) must be reduced as well as the short-term frequency fluctuation (FM noise). It is also easy to realize that a reaction stabilizing cavity works in the same way as a transmission stabilizing cavity and should cost less because of its simpler configuration. The block diagram is shown in Fig. A-1 (b). In 1970, Y. Ito et al.[16] and S. Nagano et al.[27] showed a remarkable improvement in FM noise and long-term frequency stability by the use of a reaction stabilizing cavity for Gunn diode and IMPATT diode oscillators, respectively. However, the mechanism of frequency stabilization needs further clarification.

In the following section, a general analysis of the oscillator circuit characteristics is presented. This is done by giving a general equivalent circuit applicable both to a transmission type and a reaction type stabilization. The effects of each circuit parameter is then analyzed and the self-locking mechanism is considered to give a complete understanding of the cavity stabilized negative resistance diode



(a) TRANSMISSION STABILIZING CAVITY



(b) REACTION STABILIZING CAVITY

Figure A-1 Block Diagram of Cavity Stabilized Negative Resistance Diode Oscillator

oscillator circuits. In Section A-3, the effects of environmental conditions on the stability are analyzed. Application to low-cost oscillators is discussed in Section A-4.

A-2 Operation Mechanism of Cavity Stabilized Negative Resistance Diode Oscillator

A-2-1 Load Characteristics

Figure A-2 represents the equivalent circuit both for the transmission and reaction cavity stabilized oscillators, where Q_o is smaller and G_s is bigger for the transmission type than for the reaction type because the load is included in Y_s in the former case.

The definition of unloaded $Q;Q_o$, loaded $Q;Q_L$, external $Q;Q_e$ and coupling parameter K, is as follows:

$$Q_o = \omega_o C_s / G_s$$

$$Q_L = \omega_o C_s / (1 + G_s)$$

$$Q_e = \omega_o C_s$$

$$K = Q_o / Q_e = 1/G_s$$

where

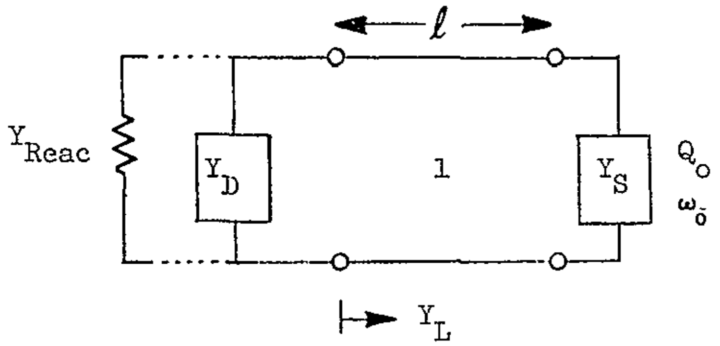
$$1/Q_L = 1/Q_o + 1/Q_e$$

$$Q_o = Q_L \, (1 + K)$$

The admittance of the stabilizing cavity viewed from the main cavity Y_s is expressed as

$$Y_s = G_s + jB_s$$

$$B_s = Q_e(\omega/\omega_o - \omega_o/\omega)$$



Y_D:NORMALIZED EQUIVALENT ADMITTANCE OF DIODE INCLUDING ADMITTANCE OF PACKAGE AND POST, IF ANY

Y_S:NORMALIZED EQUIVALENT ADMITTANCE OF STABILIZING CAVITY INCLUDING LOAD ADMITTANCE IN CASE OF TRANSMISSION STABILIZING CAVITY

Y_L:NORMALIZED EQUIVALENT ADMITTANCE OF LOAD VIEWED FROM THE DIODE PLANE TOWARD STABILIZING CAVITY

Y_{Reac}:NORMALIZED EQUIVALENT ADMITTANCE WHICH SHOULD BE ADDED TO Y_L IN CASE OF REACTION STABILIZING CAVITY TO GET TOTAL LOAD ADMITTANCE Y_L

ℓ:LENGTH OF MAIN CAVITY

Q_o:UNLOADED Q OF STABILIZING CAVITY

ω_o:RESONANT ANGULAR FREQUENCY OF STABILIZING CAVITY

G_S, C_S AND L_S:CONDUCTANCE, CAPACITANCE AND INDUCTANCE COMPONENT OF Y_S, RESPECTIVELY

Figure A-2 Equivalent Oscillator Circuit (admittances are normalized by the characteristic admittance of the main cavity transmission line)

C2

where ω is angular frequency and the admittances are normalized by the characteristic admittance of the transmission line. From the transmission equation,

$$Y_L = \frac{Y_s + \tanh \gamma \ell}{1 + Y_s \tanh \gamma \ell}$$

where γ = propagation constant

$= \alpha + j\beta$

α = attenuation constant

β = phase constant

It is assumed that the loss of the main cavity is negligible:

$$\gamma = j\beta$$

Then Y_L becomes

$$Y_L = G_L + jB_L$$

$$G_L = \frac{K^{-1}\ (1-\tan^2\beta\ell)}{(1-B_s\tan\beta\ell)^2 + K^{-2}\ \tan^2\beta\ell}$$

$$B_L = \frac{(1-B_s\tan\beta\ell)\ (B_s + \tan\beta\ell)-K^{-2}\tan\beta\ell}{(1-B_s\tan\beta\ell)^2 + K^{-2}\tan^2\beta\ell}$$

Another parameter is defined ω_m, the angular frequency where $\beta\ell = \pi$.

An example of admittance loci is shown in Figs. A-3 and A-4, where f_m = 11.4 GHz, Q_o = 20,000 and K = 25.0. When $f_o > f_m$, the admittance locus is essentially a circle and intersects with itself. The diameter of the circle decreases with increase in f_o. The arrow shows the

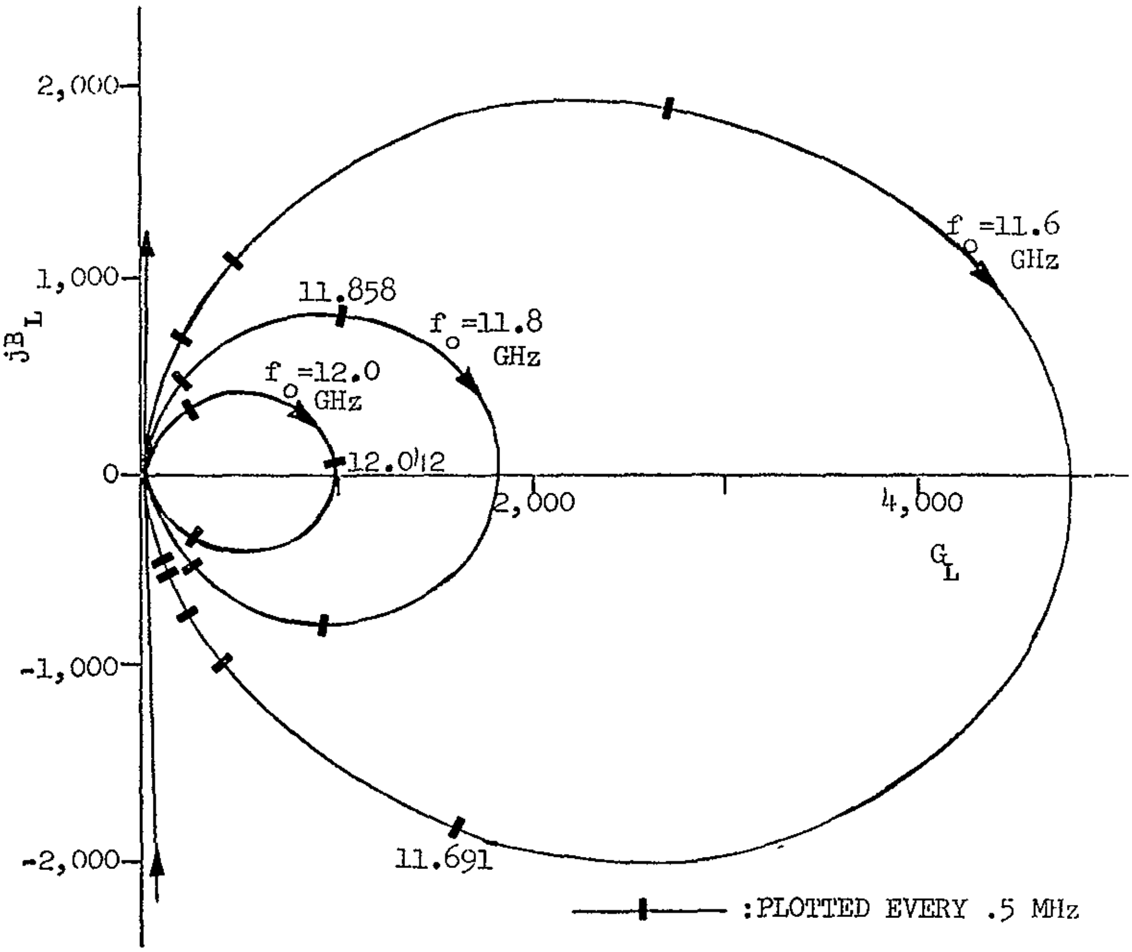


Figure A-3 (a) Normalized Admittance Y_L - effect of f_o (Q_o = 20,000; K = 25; f_m = 11.4 GHz; f_o = 11.6, 11.8 and 12.0 GHz)

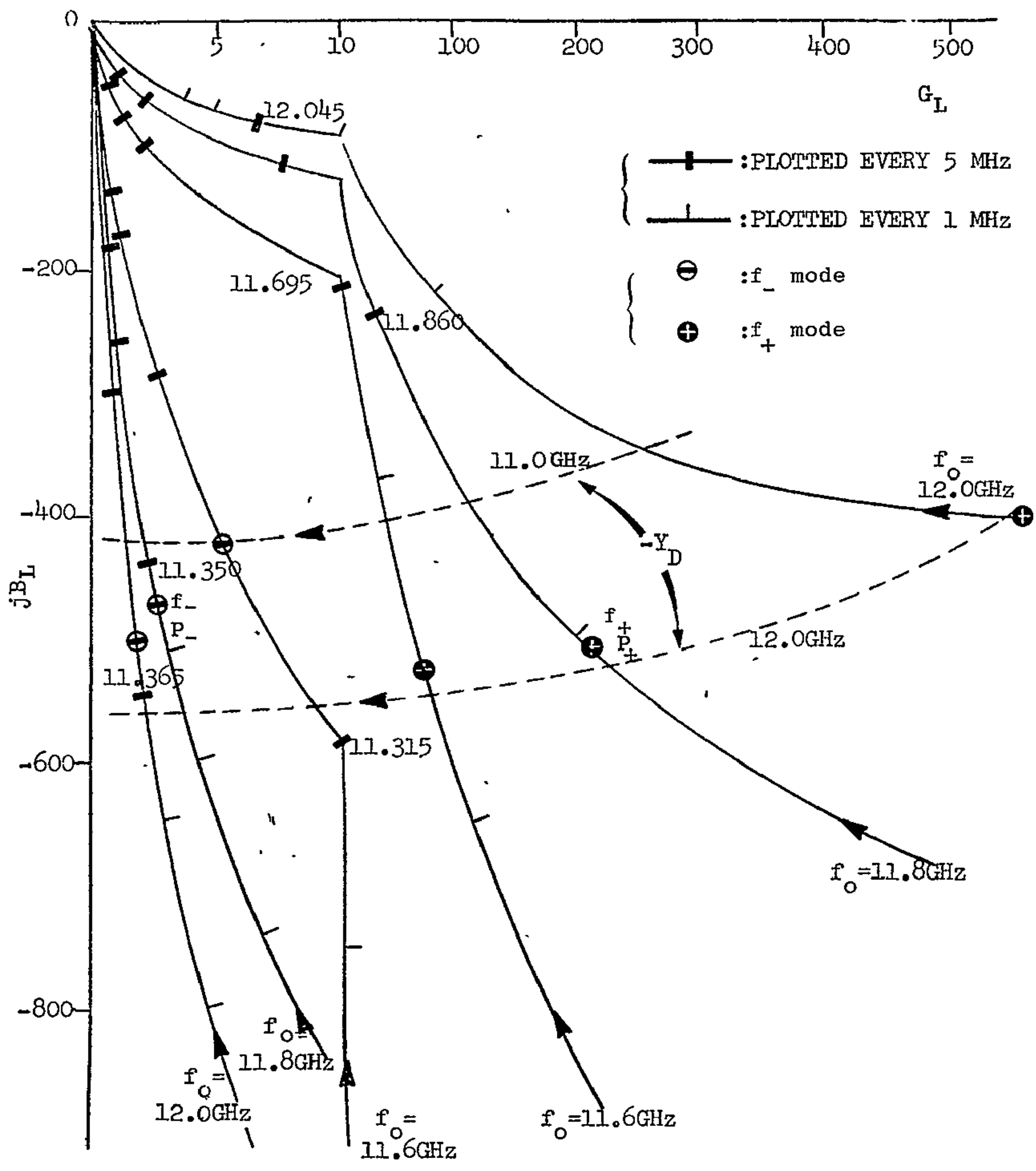


Figure A-3 (b) Enlargement of a Part of (a)

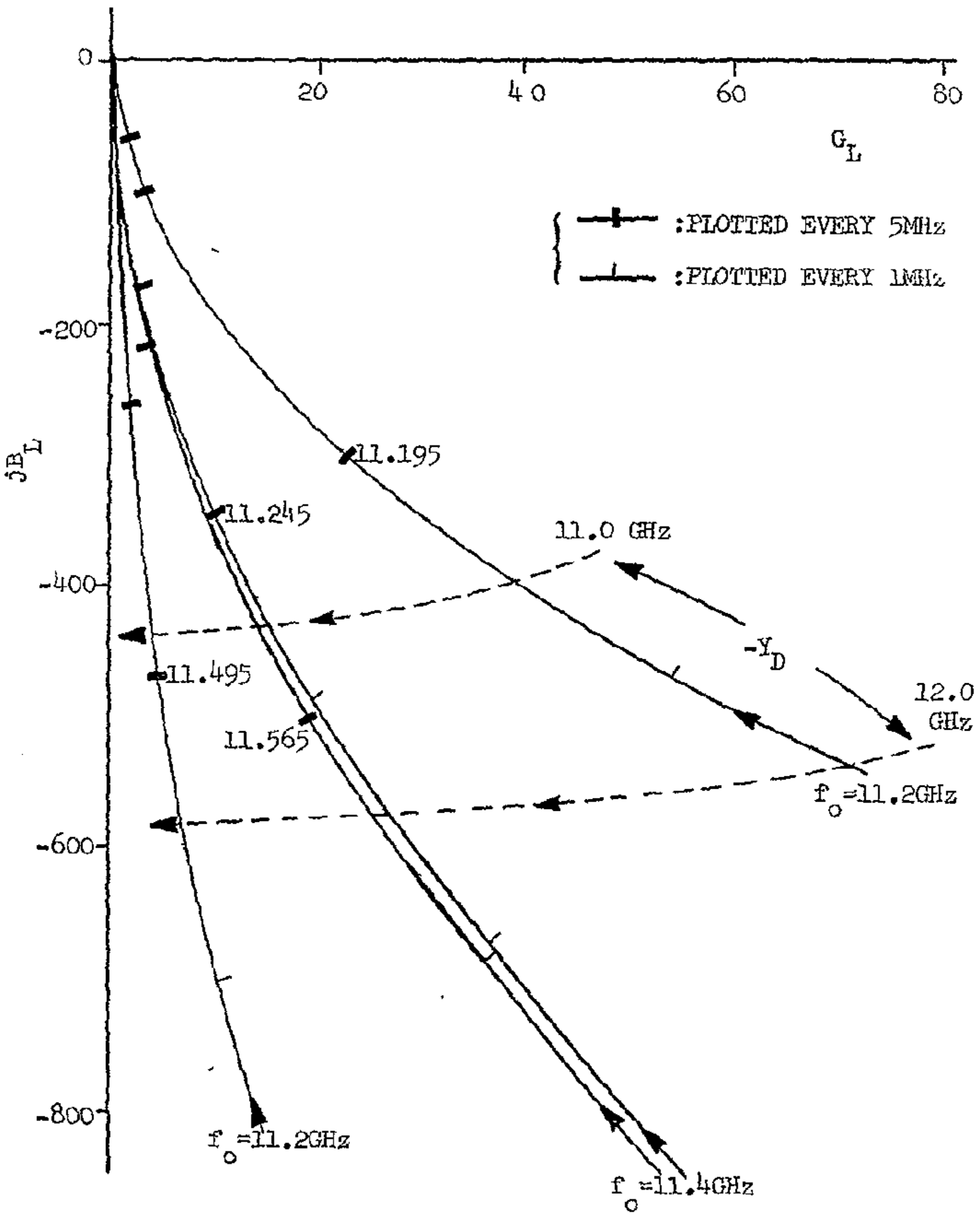


Figure A-4 Normalized Admittance (Q_o = 20,000; K = 25; f_m = 11.4 GHz; f_o = 11.2 and 11.4 GHz)

directions of the increase in frequency. At $f = f_o$, the susceptance B_L becomes zero. When $f_o \leq f_m$, there is no circle but are two loci near f_m, one below and the other above f_m.

A-2-2 Negative Admittance of Diode

The equivalent circuit of a negative resistance diode including parameters of a packaging (and a mounting post, if any) was investigated by many authors[28][32]. The circuit diagram shown in Fig. A-5 (a) given by Owens et al.[29], is useful over a fairly broad band (2 ~ 3 octaves up to 12 GHz). If we use a reduced height waveguide without a post, L_p and C_p are eliminated. A simpler equivalent circuit is shown in Fig. A-5 (b) which is valid for narrower frequency bands. With the use of the microwave strip line configuration, without a packaging, the equivalent admittance becomes almost that of the semiconductor itself.

Techniques for measuring the negative admittance of diodes were presented by the author[33][34] and others. The typical loci of the inverse of a Gunn diode admittance is shown in Fig. A-3. Those loci consist of constant-frequency contours and constant power contours. The arrow shows the direction of amplitude increase.

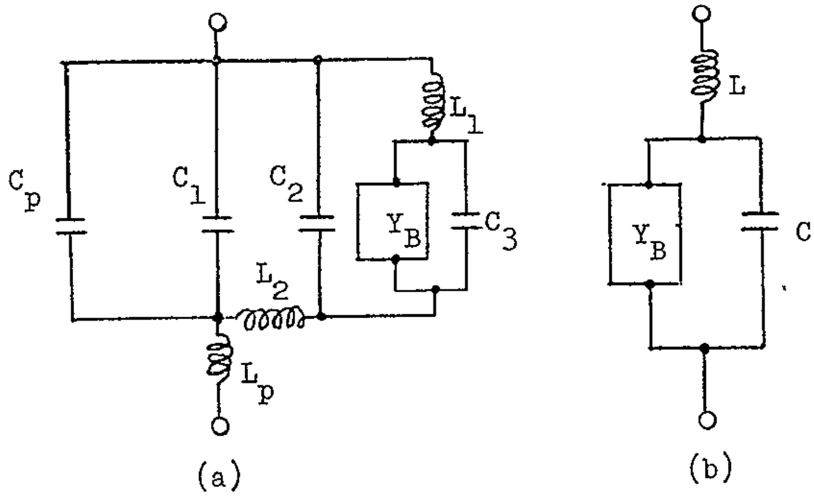


Figure A-5 Equivalent Circuit of Gunn Diode

(a) is a broad-band equivalent circuit

(b) is a simplified equivalent circuit

Typical numerical value for a standard S_4 package for

(a): $C_1 = 0.05 \pm 0.01$ pF $C_2 = 0.21 \pm 0.02$ pF

$C_3 = 0.04 \pm 0.02$ pF $L_1 = 0.23 \pm 0.005$nH

$L_2 = 0.64 \pm 0.017$nH

Y_B is the diode bulk admittance and L_p and C_p are equivalent inductance and capacitance of post, respectively.

A-2-3 Oscillating Frequency and Hysteresis

The stable oscillation occurs when $Y_D + Y_{Ltotal} = 0$. Where

$Y_{Ltotal} = Y_L$ for transmission cavity stabilization

$= Y_L + Y_{Reac}$ for reaction cavity stabilization

Let f_H be the frequency where the $(-Y_D)$ locus contacts with the Y_{Ltotal} circle at one point. When $f_H > f_o > f_m$, there are three possible points where $-Y_D$ intersects Y_{Ltotal}. But, only two out of the three are stable oscillating points[35]. An example is shown in Fig. A-3(b) where the two stable points are indicated.

Let f_- and f_+ be the lower and the higher oscillating frequency and P_- and P_+ be the corresponding oscillating power which is read from Fig. A-3. The oscillators operate usually at one mode and not at several modes at the same time. Now consideration is given to which mode, the f- or f+ is selected.

Although the f_- mode is the higher power mode $(P_- > P_+)$, the dynamic Q given by:

$$Q_{dynamic} = \frac{1}{2} \frac{dB_L}{d\omega}$$

of the f_+ mode is much higher than that of the f_- mode as we see in Figs. A-3(a) and (b). Hence the f_+ mode is much more stable than the other and the oscillation can be locked to the f_+ mode. This is interpreted as a kind of self-locking mechanism, using an analogy from the injection locking theory.

The locking range of a noisy oscillator to a stable signal source (injected stable signal) is given by[36]

$$\frac{\Delta \bar{f}_{max}}{f_{free}} = \frac{1}{Q_{ext}^{2}} \frac{P_s}{P_o}$$

where f_{free} is the free-running oscillating frequency, Q_{ext} is the external Q of the oscillator, P_s is the signal power, P_o is the output power of the free-running oscillator and Δf_{max} is the maximum frequency difference between the free-running frequency and the synchronizing frequency when the oscillator is locked.

By the analogy, one can interpret the self-locking mechanism in this case as follows. Assume that the dynamic Q of the f_+ mode is higher than that of the f_- mode. It is reasonable to consider that the oscillator is oscillating at the f_- mode with the output power P_- and that mode jumping occurs from the f_- mode to the f_+ mode under some condition such as by triggering from a noise component of the f_- mode, because the f_- mode is a higher power mode. Letting the instantaneous noise power at the frequency f_+ be P_{noise}, the locking range is then given by

$$\left(\frac{\Delta f_{max}}{f_-}\right)^2 = \frac{1}{Q_{ext}^{2}} \left(\frac{P_{noise}}{P_-}\right)$$

Let $f_M = f_- + \Delta f_{max}$

When $f_m < f_o < f_M$, the oscillator is locked at the f_+ mode but the oscillating power is not P_+ but somewhat smaller in value than P_+ (free-running power), which is verified by experimental results[25],[27]. When $f_H > f_o > f_M$, which

mode is selected is not uniquely determined but depends on its history. Hence, a hysteresis occurs.

Next we consider the case, $f_o > f_H$. As is seen from Fig. A-3, there is only one mode (f_- mode) for the oscillating mode.

Finally, when $f_o > f_m$, there are always two oscillating modes, the f_- mode and the f_+ mode, as is seen from Fig. A-4. In this case, the f_+ mode is the dominant mode because it is the higher power and more stable mode, although only slightly, than the other mode.

This discussion indicates that the oscillator shows a hysteresis phenomena as illustrated in Fig. A-6, which was reported in papers[26],[27] and verified by these experiments.

A-2-4 Effects of Circuit Parameter Change

Fig. A-3 (a) shows that more stable oscillation can be obtained with higher f_o. Figs. A-7 and A-8 show the effect of Q_o and K, respectively. Hence, in order to get more stable oscillation, that is, in order to get a higher f_M, the following is desirable:

(1) higher Q_o

(2) higher coupling factor K

Figure A-9 shows the effect of the variation of f_m. +1 and -1% variation of f_m causes about +0.12 and -0.09% variation of the oscillating frequency in this case, where Q_o = 20,000, K = 25, f_o = 11.8 GHz and f_m = 11.4 GHz. Hence, the stability factor $S = \Delta f_{osc} / \Delta f_m$ is about 7 to 10 in this case. When f_o = 12.0 GHz, S becomes more than 50, which means the frequency stability of the oscillator is almost determined by the stability of the resonant frequency of the stabilizing cavity, for all practical purposes.

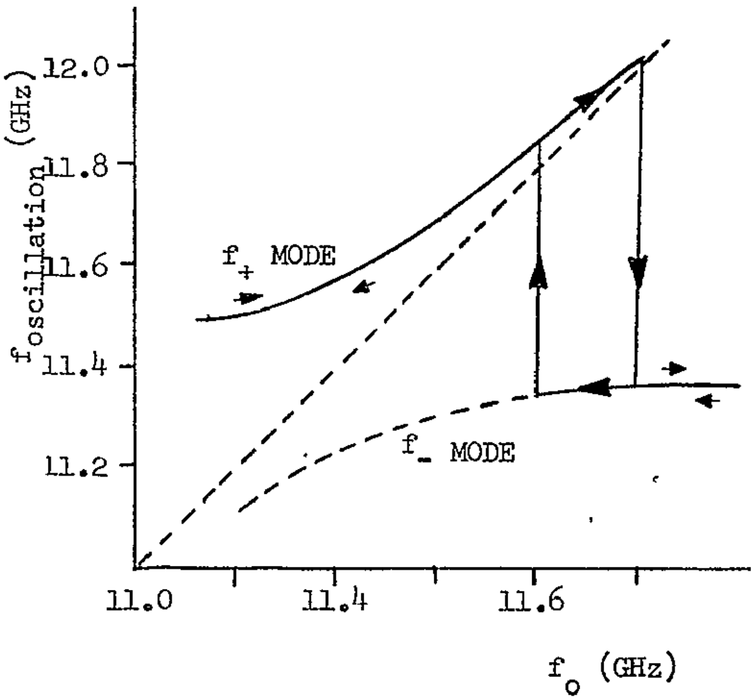


Figure A-6 Hysteresis

(f_m = 11.4 GHz; Q_o = 20,000, K = 25)

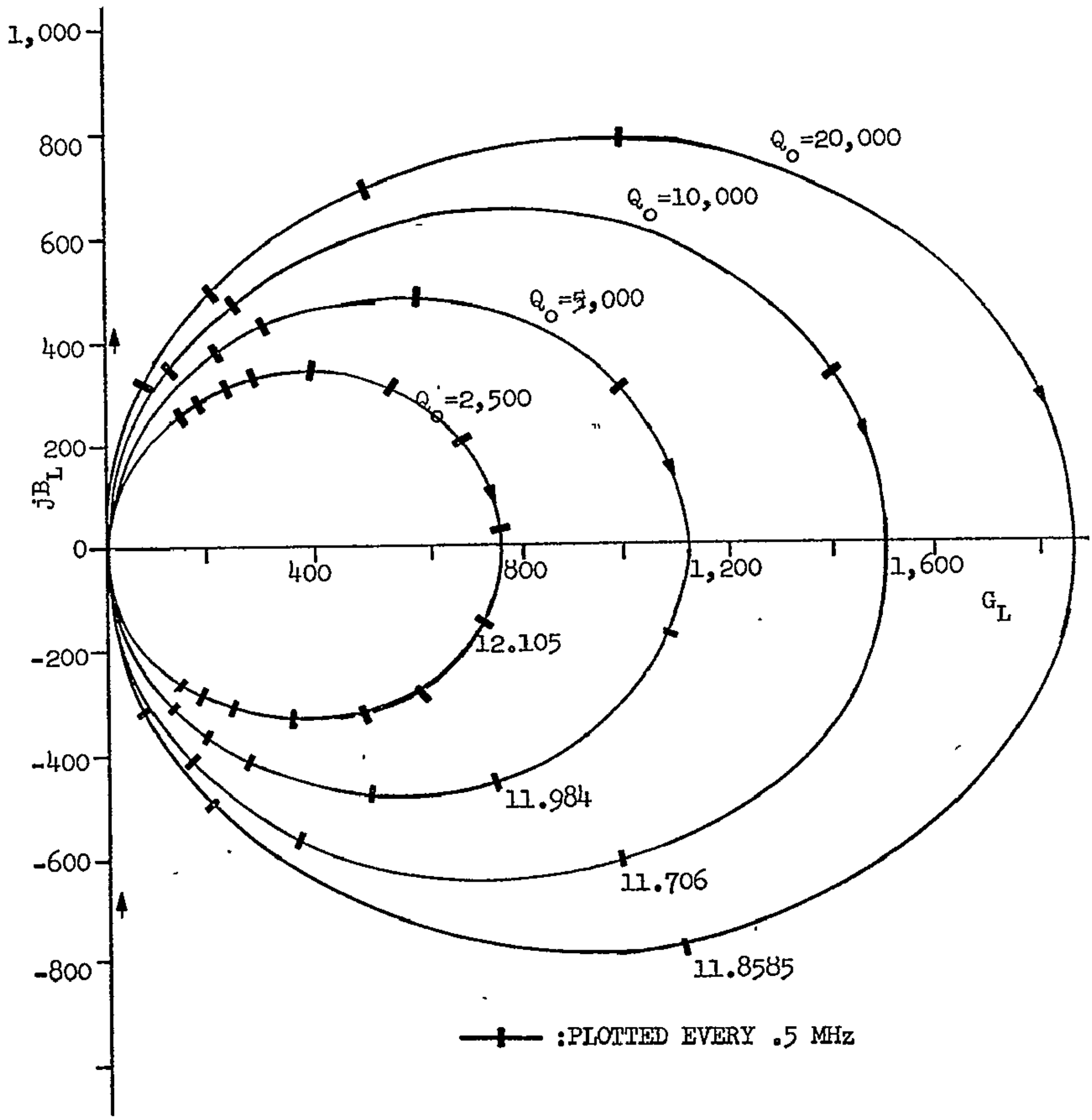


Figure A-7 Effect of Q_o on Y_L ($K = 25$; $f_m = 11.4$ GHz; $f_o = 11.8$ GHz; Q_o = 20,000, 10,000, 5,000 and 2,500).

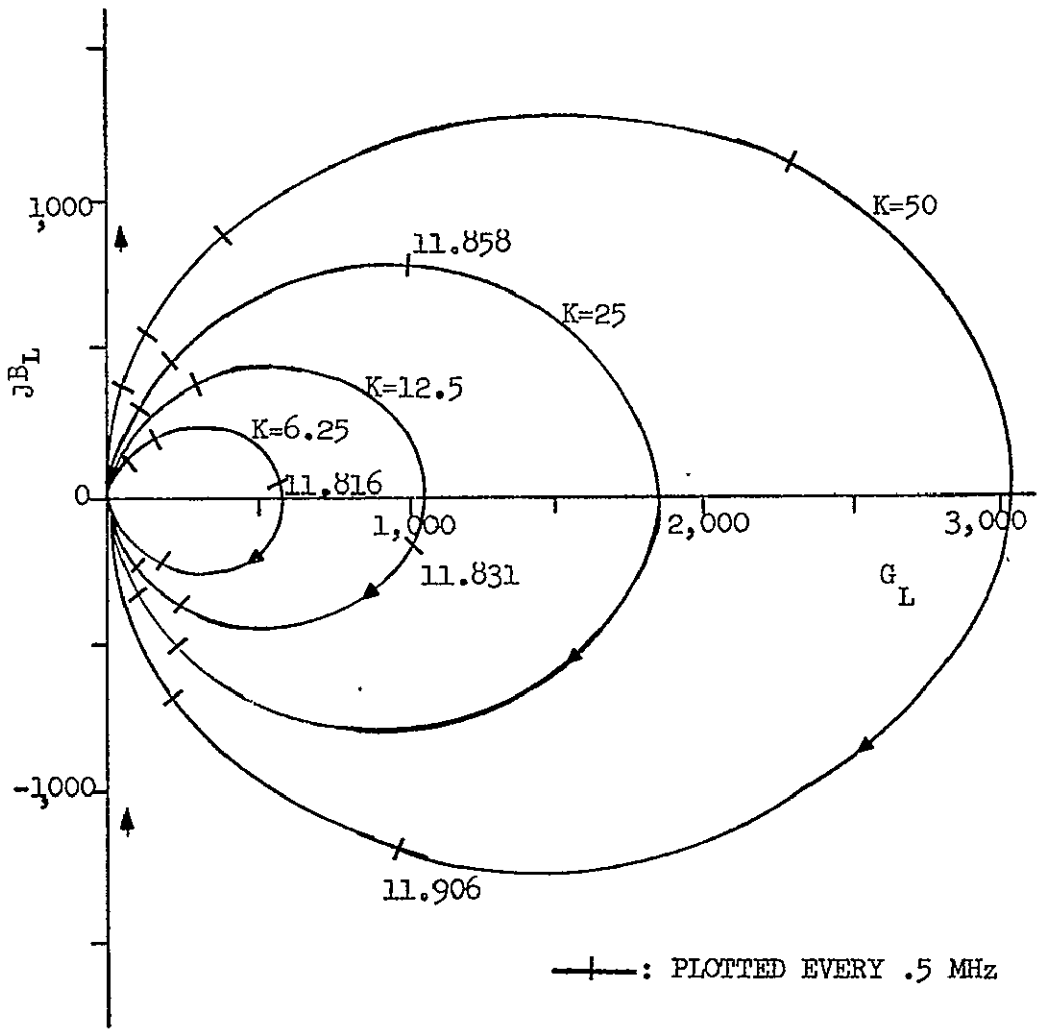


Figure A-8 Effect of K on Y_L (Q_o = 20,000; f_m = 11.4 GHz; f_o = 11.8 GHz; K = 50, 25, 12.5 and 6.25).

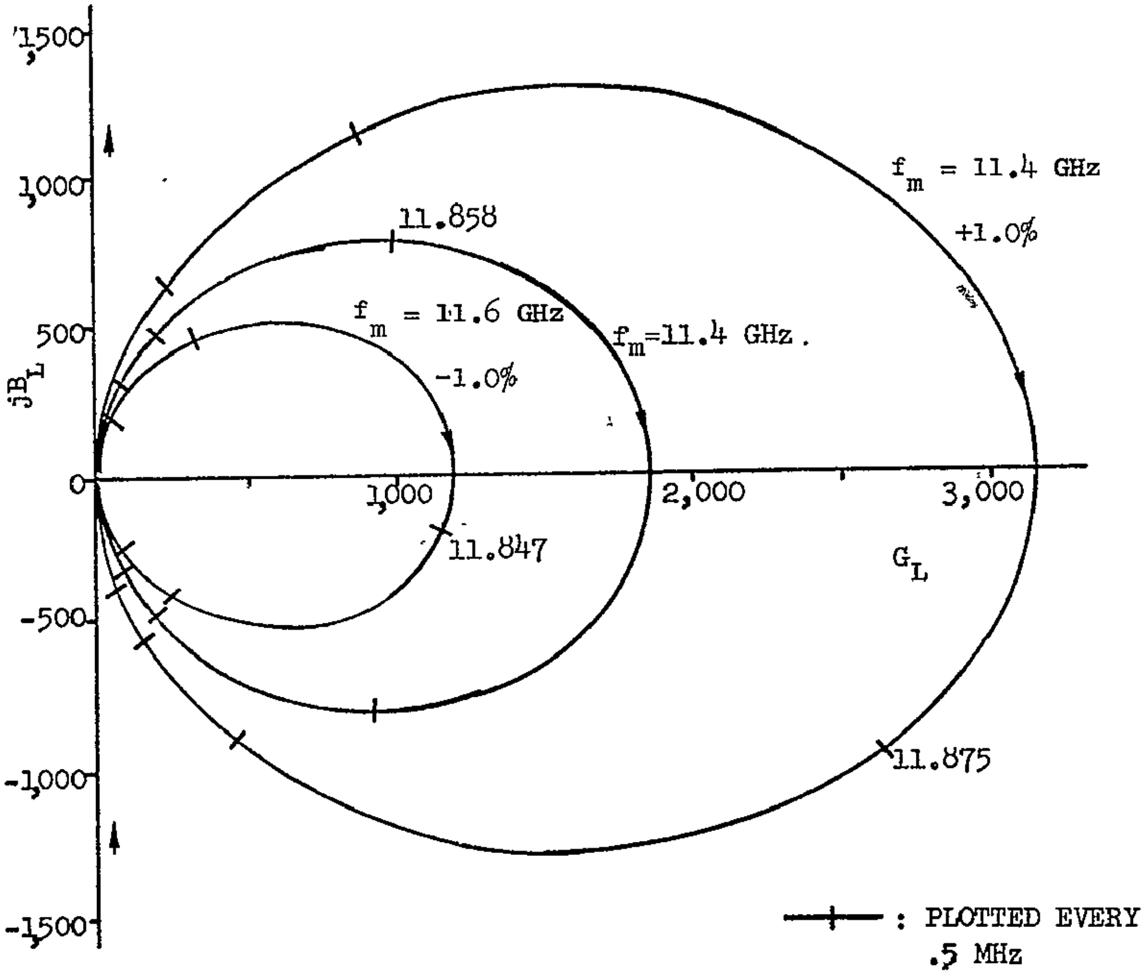


Figure A-9 Effect of f_m on Y_L (Q_o = 20,000; K = 25; f_o = 11.8 GHz; f_m = 11.4 GHz and 11.4 GHz $\pm$ 1%).

A-3 Frequency Variation Due to Environmental Condition

A-3-1 Resonant Frequency Variation of Microwave Cavities

Changes of environmental conditions cause deviation of the oscillating frequency or sometimes stops oscillation. Therefore, the effects of the following need to be known:

(1) ambient temperature change and

(2) bias voltage change

on the diode negative admittance and the effects of:

(3) thermal expansion of the cavity material

(4) dielectric constant change caused by humidity, air pressure and temperature changes, in case air is used as dielectric material inside the cavity

on the resonant frequencies of the main and stabilizing cavities.

The effects of the temperature change and the bias voltage change on the diode negative admittance have to be experimentally examined. Selecting bias voltage properly is really important for this type of oscillator[20].

The resonant frequency of a cavity is determined by the dimension of the cavity and the dielectric constant in the cavity. Assuming that the thermal expansion rate of the cavity metal and the dielectric constant in the cavity is homogeneous in every direction. Then the frequency stability is given by

$$\frac{\Delta f}{f} = -\frac{\Delta D}{D} - \frac{1}{2} \cdot \frac{\Delta \epsilon}{\epsilon}$$

$$= -\alpha T - \frac{1}{2} \cdot \frac{1}{\epsilon} \cdot \left(\frac{\partial \epsilon}{\partial T} \cdot \Delta T + \frac{\partial \epsilon}{\partial P_a} \cdot P_a + \frac{\partial \epsilon}{\partial P_w} \cdot P_w\right)$$

where

D = dimension of the cavity

ϵ = dielectric constant

T = temperature

p_a = air pressure inside the cavity

p_w = water vapor pressure inside the cavity

α = thermal expansion rate of the cavity material

and Δ represents the increment

The effect of temperature, air pressure and water vapor pressure (humidity) on dielectric constant is empirically given[37] by

$$\epsilon = 1 + 2.1 \times 10^{-4} \left(\frac{p_a}{T}\right) + 1.8 \times 10^{-4} \left(1 + \frac{5.58 \times 10^{3}}{T}\right) \cdot \frac{p_w}{T}$$

where p_a and p_w are given in mmHG and T is in ^{o}K. Then,

$$\frac{\partial \epsilon}{\partial T} = -\left\{2.1 \times 10^{-4} \frac{p_a}{T^2} + 1.8 \times 10^{-4} \cdot \left(1 + \frac{1.116 \times 10^{4}}{T}\right) \frac{p_w}{T^2}\right\}$$

$$\approx -\left\{2.1 \times 10^{-4} \frac{p_a}{\bar{T}^2} + 1.8 \times 10^{-4} \left(1 + \frac{1.116 \times 10^{4}}{\bar{T}}\right) \frac{p_w}{\bar{T}^2}\right\}$$

$$\frac{\partial \epsilon}{\partial p_a} = 2.1 \times 10^{-4} \frac{1}{T}$$

$$\approx 2.1 \times 10^{-4} \frac{1}{\bar{T}}$$

$$\frac{\partial \epsilon}{\partial p_w} = 1.8 \times 10^{-4} \cdot \left(1 + \frac{5.58 \times 10^{3}}{T}\right) \left(\frac{1}{T}\right)$$

$$\approx 1.8 \times 10^{-4} \cdot \left(1 + \frac{5.58 \times 10^{3}}{\bar{T}}\right) \left(\frac{1}{\bar{T}}\right)$$

where

$$T = T_o + \Delta T$$

$$p_a = p_{ao} + \Delta p_a$$

$$p_w = p_{wo} + \Delta p_w$$

and

$$\bar{T} = T_o + \frac{\Delta T}{2}$$

letting T_o, p_{ao} and p_{wo} be initial values when the cavity is tuned.

Simulation of frequency stability for temperature, relative humidity, air pressure and thermal expansion rate ranging from $-40^\circ C$ to $+80^\circ C$, 0% to 100%, 500 mmHG to 800 mmHG and $1.0 \times 10^{-6}/^\circ C$ (invar) to $20 \times 10^{-6}/^\circ C$ (aluminum), respectively, was done.

In all the cases, $T_o = 25^\circ C$, $P_{ao} = 760$ mmHG and the initial relative humidity is 60%.

Figure A-10 shows the frequency stability contours when invar is used and the air pressure is constant. The effect of the air pressure is negligible. 500 mmHG corresponds to the air pressure at the altitude of about 3,500 m. The effect of reduced air pressure compensates for the effect of temperature rise and humidity rise. The effect of increase in the air pressure to 800 mmHG is small enough to neglect.

A-3-2 Range of Environmental Conditions to be Considered for Practical Cases

If the oscillator is installed outdoors, it must operate over a very severe environmental condition. Although the ambient temperature varies between $-40^\circ C$ and $+50^\circ C$ (the minimum and maximum temperature

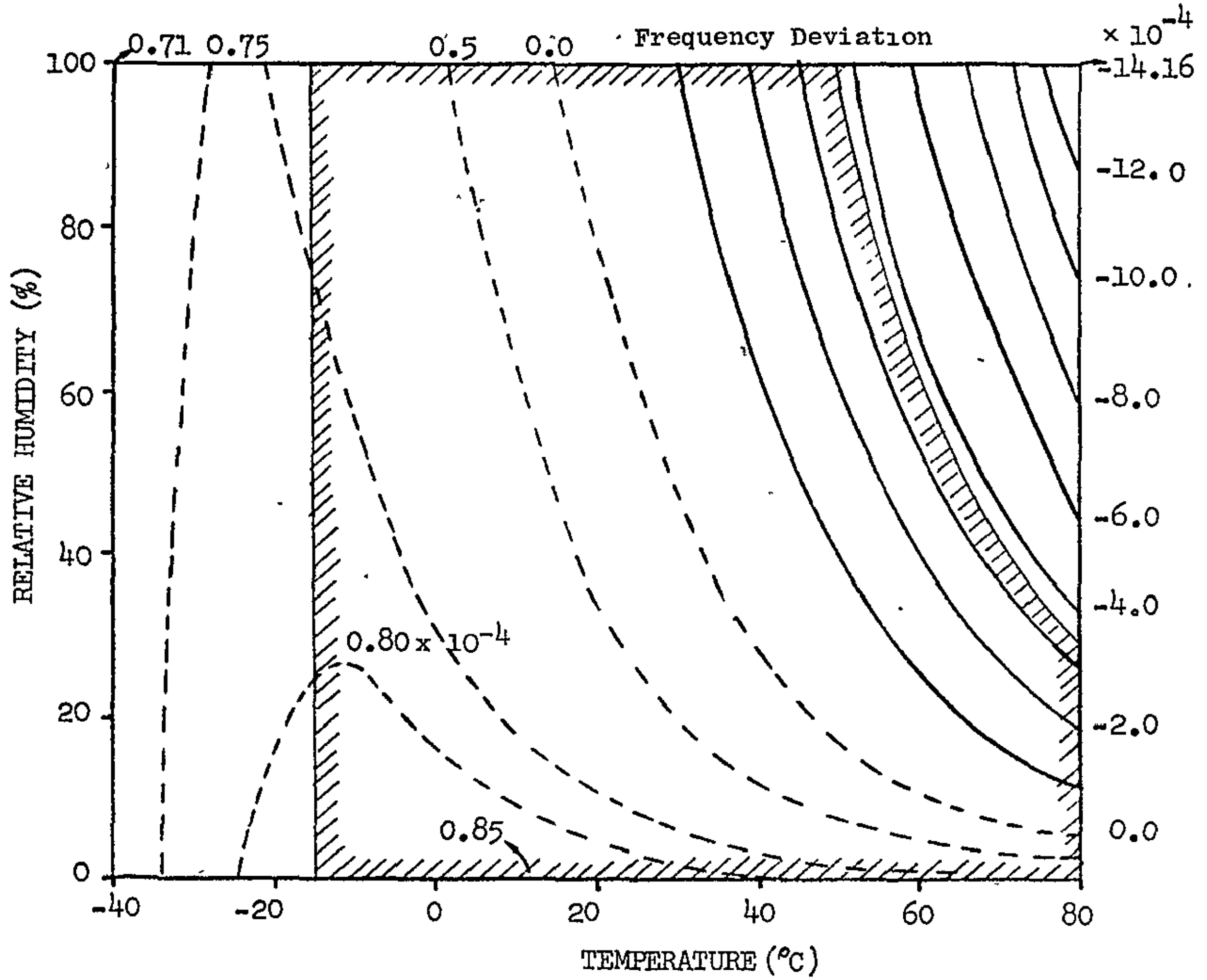


Figure A-10 Normalized Frequency Deviation Contours of the Invar Cavity (Thermal expansion rate = 1.0 x $10^{-6}/^{o}C$ and air pressure = 700 mmHG. Broad lines show actual region to be considered)

in big cities in the U.S.A. in 1969), the temperature of the oscillator cavity is always somewhat higher, because the cavity is heated by the consumed power of a Gunn diode and also by the sun in the daytime. Therefore, it is reasonable to consider $-15^\circ C$ to $+80^\circ C$ for the full temperature range. The relative humidity is sometimes below 10% in the desert and almost 100% during wet days, so that we must consider the full range (0% to 100%) of relative humidity. However, although the cavity temperature sometimes goes over $50^\circ C$, the humidity in the cavity never exceeds the saturated vapor pressure at $50^\circ C$.

These considerations give a bound on the expected frequency deviation, as shown by the solid line in Fig. A-10. Table A-1 summarizes the frequency stability for various cavity materials for this environmental condition.

A-3-3 Effect of Hermetic Sealing

As is shown in Table A-1, the frequency stability provided by an unsealed invar cavity is still relatively large because of the effect of humidity. When the cavity is sealed with dry air or nitrogen gas, Boyle's law holds for the air inside the sealed cavity and

$$\frac{p_a}{T} = \text{constant} = \frac{p_{ao}}{T_o}$$

$$\frac{p_w}{T} = \text{constant} = \frac{p_{wo}}{T_o}$$

Then

$$\frac{\partial \epsilon}{\partial p_a} = \frac{\partial \epsilon}{\partial p_w} = 0$$

Table A-1

Frequency Stability of the Microwave Cavity

(ambient temperature: -40°C~+80°
ambient humidity: 0%~100%
air pressure: up to 800 mmHG
relative humidity in sealed cavity: 5% at 25°C)

Cavity Material	Coefficient of Expansion ($10^{-6}/°C$)	Unsealed Cavity		Sealed Cavity	
		Stability (10^{-4})	Deviation at 12 GHz (MHz)	Stability (10^{-4})	Deviation at 12 GHz (MHz)
Invar	1.0	+0.8 -3.6	+0.96 -4.32	+0.4 -0.6	+0.48 -0.72
"	2.0	+1.20 -4.5	+1.44 -5.4	+0.8 -1.1	+0.96 -1.44
Thermally compensated aluminum	5.0	+2.5 -6.0	+3.0 -7.2	+2.0 -2.75	+2.4 -3.13
"	10.0	+4.5 -8.6	+5.4 -9.3	+4.0 -5.5	+4.8 -6.35
aluminum	20.0	+8.5 -14.0	+10.2 -16.8	+8.0 -11.0	+9.6 -13.2

Hence,

$$\frac{\Delta f}{f} = -\alpha\Delta T - \frac{1}{2} \cdot \frac{1}{\epsilon} \frac{\partial\epsilon}{\partial T}\Delta$$

where

$$\epsilon = 1 + 2.1 \times 10^{-4} \cdot \frac{p_{ao}}{T_o} + 1.8 \times 10^{-4} \cdot \left(1 + \frac{5.58 \times 10^3}{\bar{T}}\right) \cdot \frac{p_{wo}}{T_o}$$

$$\frac{\partial\epsilon}{\partial T} = -1.0 \cdot \frac{1}{\bar{T}} \cdot \frac{p_{wo}}{T_o}$$

Figure A-11 shows the frequency stability when the invar cavity is hermetically sealed with the initial humidity when the cavity is sealed as a parameter. Table A-1 also shows the frequency stability of the sealed cavity for several cavity materials for the temperature range described in the previous section. The initial humidity is assumed less than 5% which is easily obtained. The increase in cost for sealing is about 20%.

It can be concluded that the combination of an unsealed aluminum main cavity and a hermetically sealed invar stabilizing cavity gives the best cost/performance factor for the microwave negative resistance diode oscillator. As was described in the previous section, the pulling range of the cavity stabilized Gunn oscillator is over several hundred MHz at 12 GHz, the effect of variation of the diode admittance and the main cavity frequency change (up to 20 MHz) is easily absorbed by the use of a stabilizing cavity.

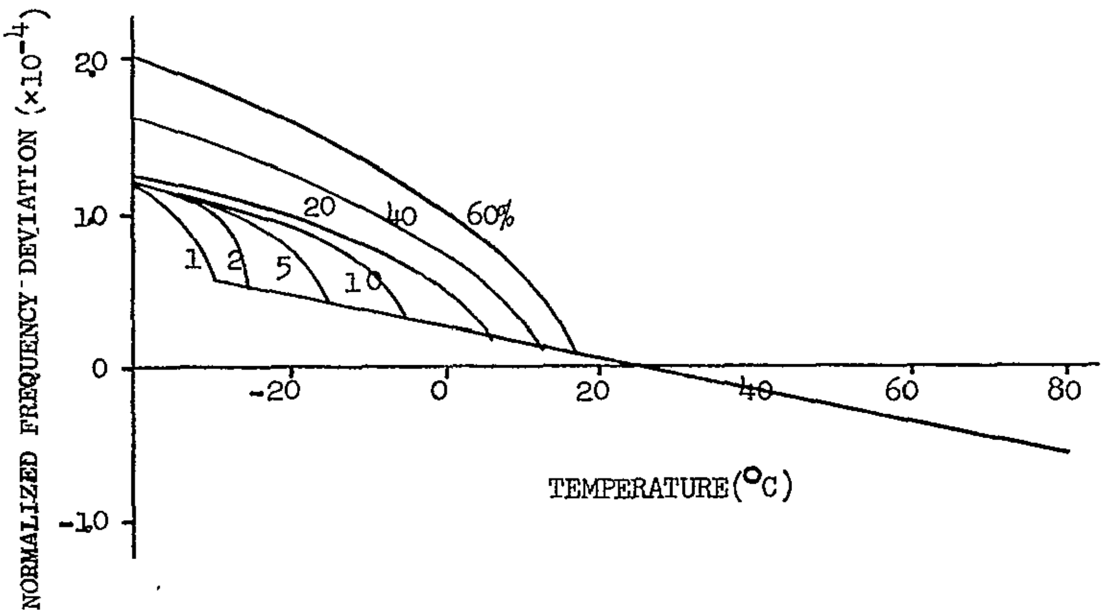


Figure A-11 Normalized Frequency Deviation of Sealed Invar Cavity
(Parameters are initial humidity. Thermal expansion rate = 1.0 x $10^{-6}/^{\circ}C$. Actual temperature region to be considered is between -15°C and 80°C.)

Conclusion

A-4 Application to High-Stability Low-Cost Oscillator

The analyzed oscillator is most suitable for high-stability low-cost microwave oscillators. Circuit configuration can be either a waveguide or a microstrip line. In the former case, the stabilizing cavity is an invar cylinder cut out from a tube and attached to an invar-and-brass terminal plate on each side by screws. The invar cavity can also be cast for low-cost in case of large quantities. In the latter case, the stabilizing cavity is composed of a quartz. Cost/performance of these configurations is discussed in Chapter III.

Appendix B

Decision Analyses of Subsystems

Introduction

B-1 Signal and Power Feeding System for 12 GHz Receiver

The signal from the satellite is converted to a VHF signal and is fed to TV sets by the receiver which has two local oscillators. Because the TV set has a local oscillator, the whole system has three local oscillators. This could raise serious interference problems, without careful consideration to avoid spurious radiations.

In addition, as a Gunn oscillator consumes as much as 3 w and requires a stable and harmless DC source, the power feeding system has to be sufficient in size and stable in capability and in safety in order to fulfill the Underwriters Laboratories (UL) requirements.

B-1-1 Signal Feeding System Configuration and Alternatives

Signal frequency and modulation:
receiver input; 12.0 GHz, FM
receiver output; VHF/AM or VHF/AM-VSB

Figure B-1 shows the block diagram of the signal feeding system. There are two alternatives:

[A] One-touch system

The consumers have only to touch the tuner of the TV set in order to select their desired signal. The output signal of the receiver is directly coupled to feed to the VHF tuner in the TV set.

[B] Two-touch system

The receiver (in this case, the indoor unit) has a switch to select either a satellite signal or ordinarily broadcast VHF signals.

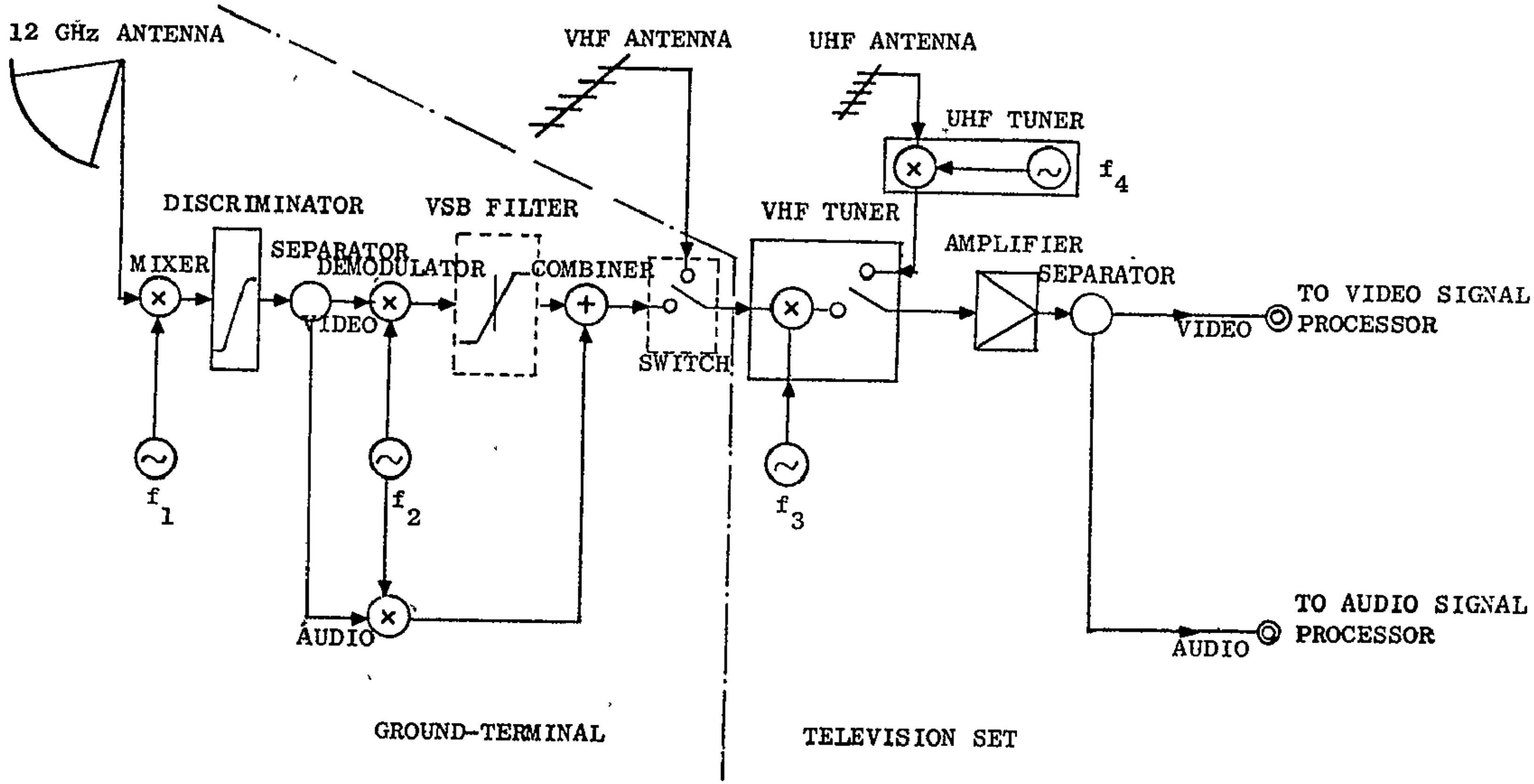


Figure B-1 Signal Flow Chart

The consumers select either one by switching the indoor unit and then selecting a channel by the VHF tuner of the TV set.

[A] is more convenient than [B] for the consumers but has serious interference problems, the four majors of which are described as follows:

(1) The lower sideband of the receiver output signal interferes with the neighboring lower channel if a VSB filter is not installed in the receiver.

(2) The beat signals between the receiver output signal and harmonics of VHF tuners local frequency (f_3 in Fig. B-1) often drop in the IF band of the TV set.

(3) The beat signals between the local frequency of remodulator (f_2) or its harmonics and the local oscillator frequency of the VHF tuner or its harmonics drop in the IF band of the TV sets.

(4) In case more than one TV signal from the satellite is broadcast in the future, the neighboring signals interfere with the neighboring channels of the TV sets.

In the author's opinion, nothing is more costly than struggling with interferences after installation. Consequently alternative [B] was chosen. This alternative needs a channel switch to be placed on or near the TV set. It is also concluded that an indoor unit is needed with a switch that can be handled by the consumer.

B-1-2 Signal Feeding Cable

The usual ribbon-type TV feeder is very lossy when it is wet (at 200 MHz) and has a short lifetime (less than three years in the districts near the sea, so that a coaxial cable is the only alternative.

B-1-3 Power Feeding System

Figure B-2 shows the block diagram of the power feeding system. Any part of the adaptor, including cables has to be isolated to fulfill the UL requirements. The DC source should be packed in the indoor unit and the DC power should be fed to the antenna unit through a pair cable to avoid noises.

B-1-4 Separation of Receiver into an Outdoor Unit and an Indoor Unit

The above considerations lead to the separation of the receiver into an outdoor unit and an indoor unit. The outdoor unit includes a feed, a local oscillator, a mixer and low-noise IF amplifiers. The indoor unit includes IF amplifiers, a limiter, a discriminator, a remodulator, a switch and a DC source. Figure B-3 shows the block diagram of the system.

B-2 Alternative Devices for the 12 GHz Local Oscillator

Microwave tubes (Klystron, Magnetron, TWT, etc.) need a higher voltage source and have a shorter life time than solid state devices. Varactor chain multiplication puts out smaller power and is much more expensive, therefore limiting the choice to Gunn diodes and IMPATT diodes.

B-2-1 Interaction Model and Outcomes

B-2-1-1 Performance Comparison

(1) Output Power Capability (continuous wave) at 12 GHz

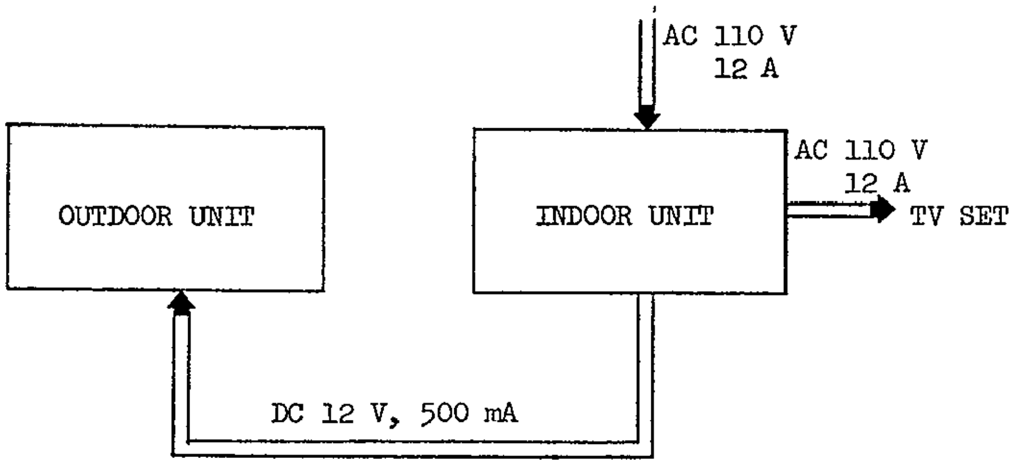


Figure B-2 Power Feeding System

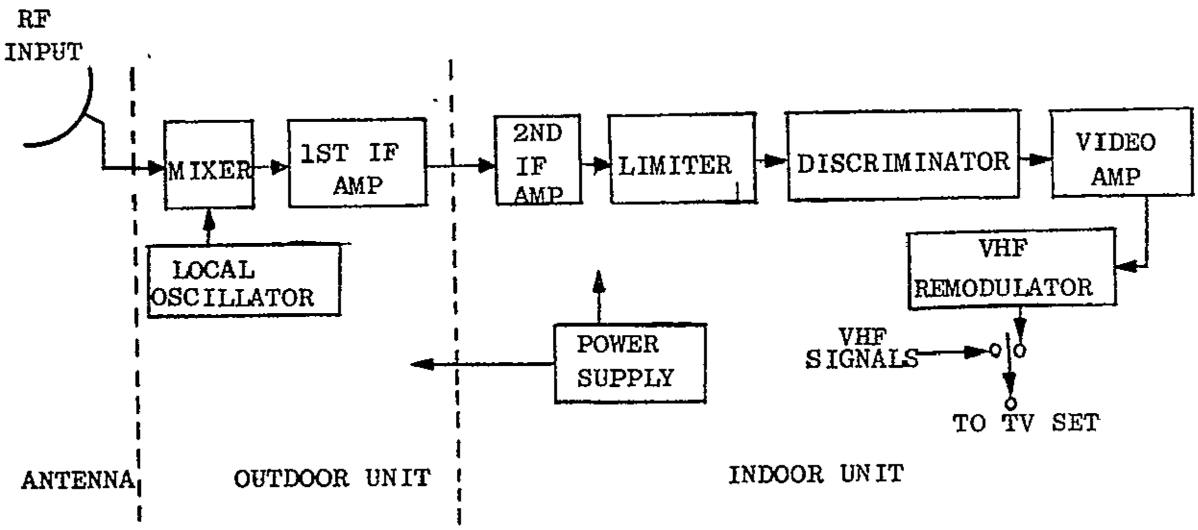


Figure B-3 Block Diagram of the Ground-Terminal

Gunn; 1 w each

IMPATT; 2-3 w each

Necessary power is less than 10 mW. IMPATTS give higher output power than Gunns[38], but this is not an advantage in our case, where Gunn diodes offer enough output power.

(2) Heat Sink and Efficiency

Diodes should be mounted on metal or some other good heat sink material. This is an important point when using a microstrip line oscillator. But there is not much difference between Gunn diodes and IMPATT diodes.

(3) Bias Voltage

Gunn diode; 6~12 V

IMPATT diode: >12 V (20~100 V, typically 60~ 90 V for X band) needs another DC source or a DC-DC converter[39].

The Gunn diode has a great advantage in that the ordinary 12 V DC supply can be used. On the other hand, the IMPATT oscillator would need another DC source or a DC-DC converter which would cost \$5~\$10 for 1,000 units[39].

(4) Oscillation Frequency

Gunn; 1~60 GHz

IMPATT; 1~150 GHz (200 MHz~340 GHz in laboratory)

Parallel resonant mode frequency (say Ku-band) is higher than series resonant mode frequency (then x-band)[40]. There is not much difference between Gunn and IMPATT.

(5) Noise Performances

a. FM Noise

General characteristics[38];

Gunn diode; -9 dB/octave

IMPATT diode; almost flat

Free-running stability (without stabilization)

Gunn: 30~50 Hz rms in 1 KHz BW, 100 KHz away from the carrier[41]

IMPATT: 70 Hz rms in 1 KHz BW, 100 KHz away from the carrier[39]

These values are comparable to those of microwave tubes.

With cavity stabilization

Gunn diode; 8 Hz rms in 1 KHz BW, 100 KHz away from the carrier[26]

IMPATT; 15 Hz rms in 1 KHz BW, 500 KHz away from the carrier[27]

b. AM Noise

Gunn; -120~130 dB/1 KHz BW, 100 KHz away from the carrier[41]

IMPATT; -133 dB/100 Hz BW, single-side-band[39]

IMPATTs are superior in the AM noise performance, but of much importance in cases like this one.

B-2-2 Cost Comparison

[Cost of Gunn diode] = K x [Cost of IMPATT diode]

Today, K = 4~10 for 100~200 mw output power[38][39]. K would be less for lower power diodes.

IMPATT diodes cost $10 each in production quantities for 100~200 mw output at X band (e.g., Hewlett-Packard 5082-0432)[39].

Future, cost of Gunns and IMPATTs is decreasing almost every month and hence, a low power Gunn diode of less than $5 would be available for 100,000 production units in a few years.

B-2-3 Value Model and Decision of Oscillator Device

Ranking of outputs of the interaction model according to the order of importance:

(1) Cost for a 12 GHz, 10 mw output oscillator is less for IMPATTs

(2) DC bias supply voltage is 12 v for Gunns and about 80 v for IMPATTs

(3) There is no distinct difference between Gunn and IMPATT in:
  a. output power
  b. frequency stability
  c. lifetime
  d. oscillation frequency
  e. circuit configuration

(4) Generally, Gunn diodes are less noisy than IMPATT diodes.

Although an IMPATT diode costs less than a Gunn diode, the former needs a higher DC supply voltage than 12 v which adds an extra cost and circuit complexity. Therefore, Gunn diode oscillators were used.

B-3 The 12 GHz Mixer

Alternatives:

a. waveguide orthomode balanced mixer
b. stripline balanced mixer
c. microstrip balanced mixer
d. unbalanced mixer for the three circuit configurations

Value Model:

Noise figure and cost are main concerns. Unbalanced mixers are unlikely to be used because of 3 dB more noise figure than balanced mixers.

Decision:

A decision must be made by cost analysis (refer to Chapter III).

B-4 Polarization

Either a linearly polarized plane wave (LP) or a circularly polarized plane wave (CP) can be used for signals transmitted from the satellite to the ground stations. Table B-1 shows the comparison.

LP is a bit cheaper and has enough performance. But for the future diversity, the design which works both for LP and CP is desirable. Hence, a circular waveguide feed should be selected.

Table B-1

Comparison Between LP and CP

	LP	CP
Rotation due to ionosphere	negligible (less than 1 dB)	does not matter
Cross-polarization	small enough (less than -25 dB)	small enough (much less than -25 dB)
Function needed to the feed	either rectangular or circular waveguide	circular waveguide and a polarizer
Adjustment needed in the field	adjustment of the feed angle around the antenna axis in addition to the antenna pointing	antenna pointing only
Cost Manufacturing Assembly	adjustment of the feed rotation	cost of filter

Appendix C

Design of the 12 GHz Receiver Subsystems

C-1 The 12 GHz Cavity Stabilized Gunn Oscillator Design

C-1-1 Oscillator Configuration

A Gunn diode oscillator stabilized by a reaction type invar cavity is to be designed. Figure C-1 shows the overview. A Gunn diode is mounted in a reduced-height TE_{101} mode cavity whose resonant frequency is about 500 MHz below the desired oscillating frequency (11.88 GHz in case f_{IF} is 120 MHz). The main cavity (reduced height rectangular waveguide) is butted against a circular iris in the side of the TE_{011} mode rigid cylindrical cavity so that the broad dimension of the waveguide is parallel to the axis of the cylinder. This TE_{011} mode cavity has high Q and low thermal expansion rate in order to stabilize the oscillating frequency against the ambient temperature rise.

C-1-2 Decision of Main Cavity

Uncontrolled variables (design criteria);

oscillating frequency: 11.88 GHz
tuning range: +100, -600 MHz
material: cast aluminum

Decision variables;

waveguide: rigid rectangular or right cylindrical

mode: $TE_{\ell mn}$ or $TM_{\ell mn}$ where ℓ, m and n are integers

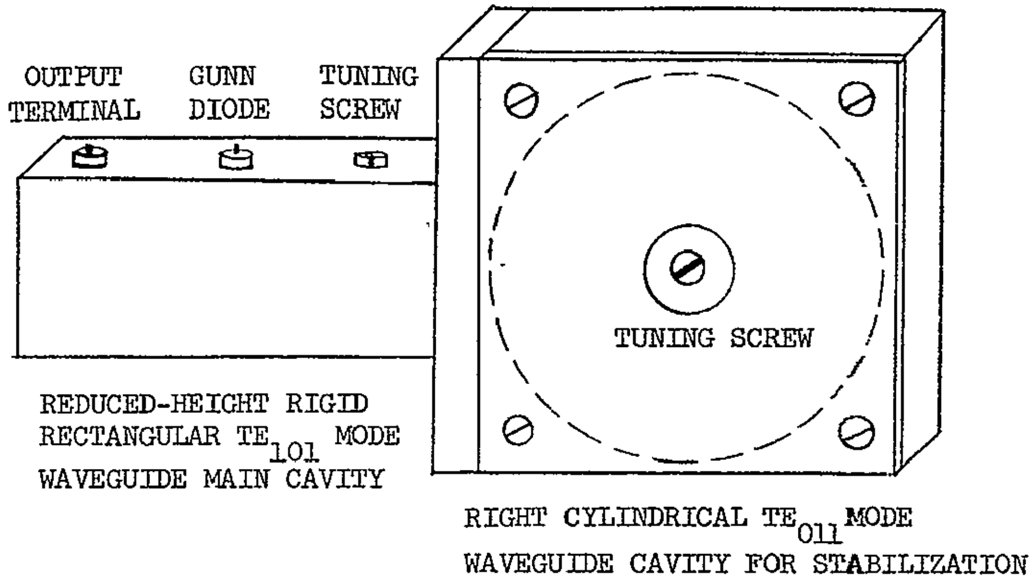


Figure C-1 Overview of Local Oscillator

Outcomes;

- cost
- ease of die casting
- tunability
- spurious mode suppression
- ease of diode mounting
- (Q is not a main concern)

Interaction model:

Table C-1 is a comparison of three typical modes in view of outcomes. It can be concluded, from the considerations given, that a reduced height TE_{101} mode cavity should be used for the main cavity.

Table C-1

COMPARISON OF THREE TYPICAL MODES

	TE_{101} IN RECTANGULAR WG	TE_{111} IN CYLINDRICAL WG	TE_{011} IN CYLINDRICAL WG
Tuning range.	22%	19%	7%
Tuning mechanism	End plate or capacitive screw	End plate	End plate
Unloaded Q at 12 GHz. . .	8,000	13,000	25,000
Diode mounting.	Easy	Difficult	Difficult

Note: To avoid a spurious mode and spurious resonance, the post for the diode mounting has to be short. Next, a mode whose current does not flow on the side walls is preferable because the cavity can be split into two pieces which is convenient for die casting.

C-1-3 Decision of Type and Mode of Stabilizing Cavity

Uncontrolled variables;

frequency: 11.88 GHz

tuning range: +100, -500 MHz

material: cast invar

Decision variables;

waveguides: (rectangle, cylinder or dome,....)

modes: $TE_{\ell mn}$ or $TM_{\ell mn}$

Outcomes;

Q

Q/volume

cost

spurious mode suppression

Interaction model

Q should be as high as possible, generating the question: what mode gives the highest Q per volume? J. P. Kinzer has shown that TE_{01} modes in a cylindrical waveguide give the minimum volume for a given Q and λ[42]. It has a simple shape both for casting and machining so that TE_{011} mode of a rigid cylindrical waveguide cavity should be chosen.

Cavity Design

(1) TE_{101} mode main cavity

For $TE_{\ell mn}$ mode $$\lambda = \frac{2}{\left\{\left(\frac{\ell}{A}\right)^2 + \left(\frac{m}{B}\right)^2 + \left(\frac{n}{C}\right)^2\right\}^{1/2}} \quad \text{(cm)}$$

where

λ: resonant wavelength

A, B, and C are dimensions of the cavity; width, height, and length, respectively.

For TE_{101} mode

$$\lambda = \frac{2}{\left[\left(\frac{1}{A}\right)^2 + \left(\frac{1}{C}\right)^2\right]^{1/2}} \qquad \text{(cm)}$$

Actually, C is the length of the waveguide between the diode and the aperture.

Resonant frequency is independent of B. The longer B is, the more series inductance there is due to the post. Hence, B should be as short as the length of the diode packaging, which is about .5 cm if the S4 package is used. Using a Fairchild's Gunn flange; A = 2.0 cm, B = 1.65 B = 1.65 cm, C = .5 cm by considering the ease of tuning by a screw. Then the unloaded Q is calculated by

$$Q\,\frac{\delta}{\lambda} = \frac{B}{2} \cdot \frac{(A^2 + C^2)^{\frac{3}{2}}}{A^3(C + 2B) + C^3(A + 2B)} \qquad \text{for } TE_{101}$$

where δ is skin depth.

(2) TE_{011} mode stabilizing cavity

The resonant frequency in a right cylindrical cavity is given by a mode chart[43]. The tunability of the cavity is performed by an end screw so that the lowest frequency f_L = 11.2 GHz.

From the mode chart, TE_{311} which is the nearest spurious mode must be avoided.

Hence,

$$\left(\frac{D}{L}\right)^2 \geq 2.0$$

$$(f_L \cdot D)^2 = 17.5 \times 10^{20}$$

which gives

diameter: $D = 3.72$ cm

length: $L = 2.63$ cm

highest frequency: $f_H = 12.2$ GHz

The inside surface of the stabilizing cavity must be silver plated in order to get high Q.

There are two methods to couple the stabilizing cavity to a rectangular waveguide main cavity[37]. (Method A) the main cavity is butted against a circular iris in the side of the cylindrical cavity so that the broad dimension of the main cavity is parallel to the axis of the cylinder. (Method B) the cylindrical cavity and the main cavity are milled and soldered together so that the circular iris is centered on the broad face of the waveguide main cavity. The axis of the cylinder is parallel to the broad dimension of the waveguide.

(Method A) was chosen because of the ease of fabrication and efficiency of space.

C-2 The Design of the 12 GHz Integrated Circuit Mixer

C-2-1 Design Criteria

A MIC singly balanced mixer was chosen out of several alternatives discussed in B-2 and Chapter III.

Design Criteria:

Type: Rat-race hybrid

Substrate: Alumina (dielectric constant $\varepsilon_o = 9.7$)

Inputs: Signal frequency = 12.0 GHz
Local oscillator frequency = 11.88 GHz

Output: Intermediate frequency = 120 MHz

Input and output impedances: 50 ohms

The principal configuration is shown in Fig. C-2. The local oscillator signal comes out of the IF ports with equal amplitude and in phase but never does come out of the signal port. The signal comes out of the IF port in equal magnitude but out of phase and never does come out of L.O. port.

C-2-2 Determination of the Width of the Microstripline

A chart of characteristic impedance is given by G. J. Wheeler[44].

For Z_o = 50 ohms -------- W/H = 0.9 -------- W = .0225" where W is the width and H is the height of a microstripline.

For Z_o = 71 ohms -------- W/H = .30 -------- W = .008" where $\varepsilon_o = 9.7$ and H = .025".

But the experiments show the necessity of some amendment of the theory[45] which gives finally;

For Z_o = 50 ohms ---------------- W = .025"

For Z_o = 71 ohms ---------------- W = .0088".

C-2-3 The Determination of Radius and Length

Experiments show that due to the fringe effect of the microstripline, the effective dielectric constant ε_{eff} is given by:

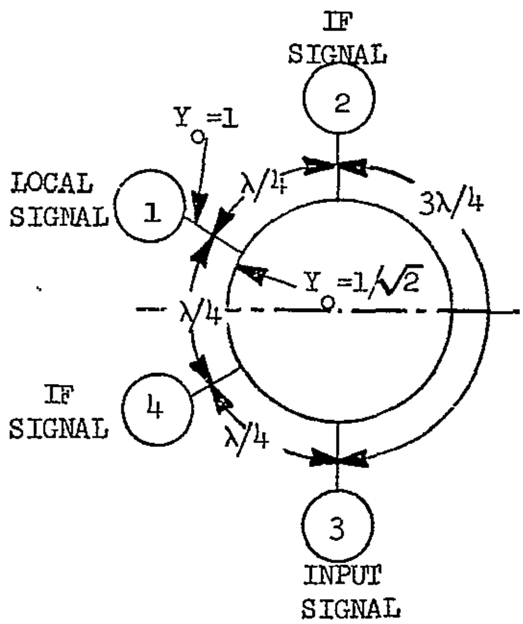


Figure C-2 Principle of Rat-Race Hybrid Mixer

$$\varepsilon_{eff} = .61\varepsilon_o$$

Hence, the needed length of the hybrid circuit (6 quarter wavelength) $L_{3/2}$ becomes:

$$L_{(3/2)\lambda} = (3/2)\lambda_o/\sqrt{\varepsilon_{eff}} = .61''$$

Now, consider the effect of the four arms on the total length. The experiments[45] show that by simply adding the width of the four ports, the results were as follows:

$$L_{total} = L_{(3/2)\lambda} + 4 = .013''$$

Figure C-3 shows the sketch of the designed mixer.

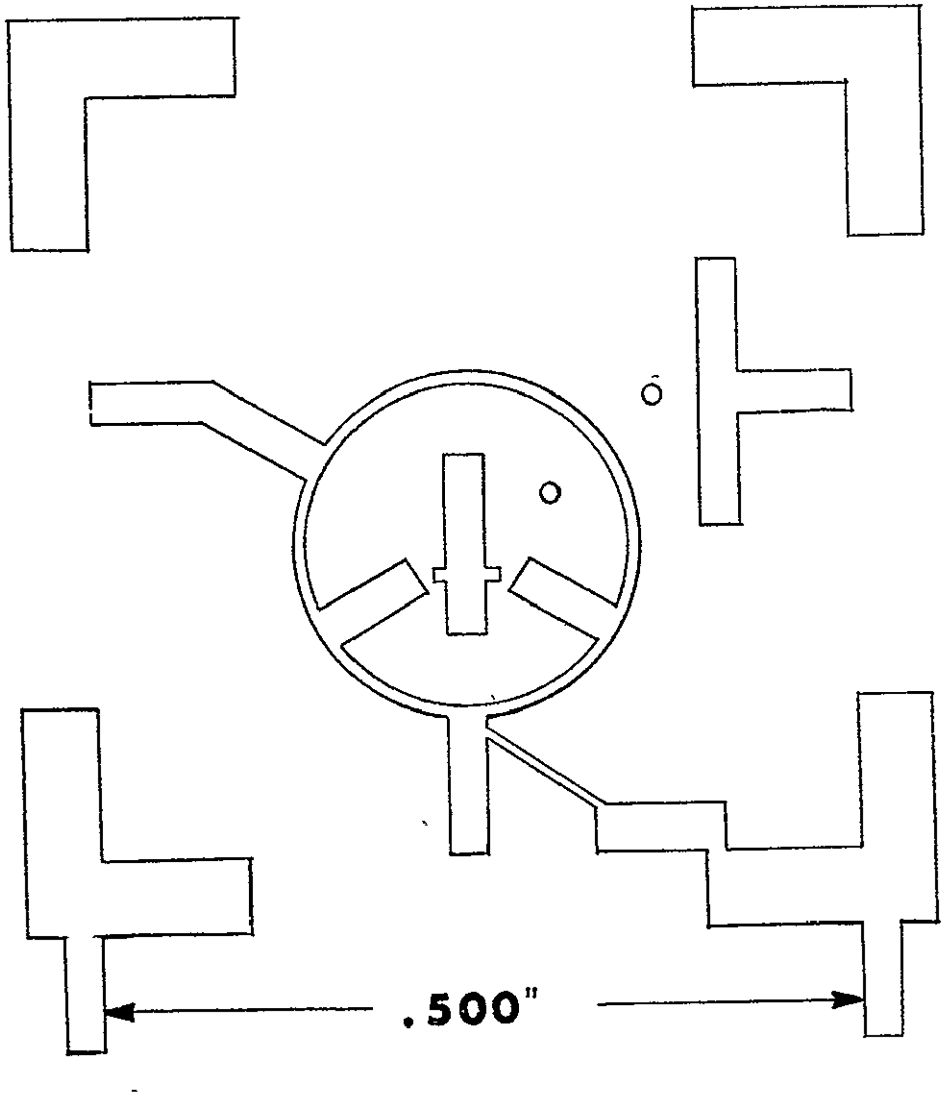


Figure C-3 Sketch of the 12 GHz MIC Mixer

C-3 The Design of the 12 GHz Receiver Packaging for Mass Production

C-3-1 Design Criteria

The following should be considered before proceeding to actual design work.

(1) The packaging should be die cast aluminum, while the stabilizing cavity is separately cast invar which is made by the use of low-pressure casting or machining.

(2) The cross-sectional area of the packaging in the direction of the satellite should be as small as possible in order to avoid the reflection of the input signal. The gain of the antenna G is given by

$$G = G_o \cdot \frac{D_a^2 - D_p^2}{D_a^2} = G_o \left(1 - \frac{D_p}{D_a} \right)^2$$

where D_a and D_p are the effective diameter of the antenna reflector and the antenna unit packaging respectively and G_o is the gain when $D_p = 0$ (see Fig. C-4). From this standpoint, a cylinder is the most desirable shape for the packaging.

(3) The whole packaging has to be fixed at the focus of the antenna, for which the cylindrical structure is again desirable for easy supporting.

(4) The packaging should protect the circuitry from the weather, especially from rain. Hence, a cap to cover the circuitry is necessary.

(5) As few die as possible are desirable because of low cost and high precision of assembly.

Using the TE_{11} mode for the feed and the TE_{101} mode for the local oscillator main cavity, the whole packaging can be split into two pieces.

(6) A symmetrical structure is desirable because of the elimination of extra die.

(7) The first IF amplifier circuitry has to be packaged in it.

C-3-2 Design

These considerations lead to the symmetrical two piece structure shown in Fig. 3-3 where the IF circuitry is packed in the top and/or bottom room. The circuitry is covered by a cylindrical cap resulting in a structure shown in Fig. 3-4. The circuitry is completely

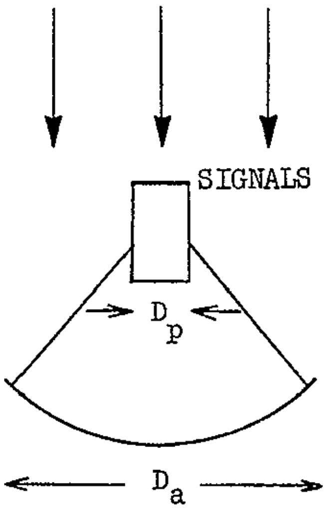


Figure C-4 Blockage of Signals by Packaging

protected from rain. The final dimension is about 7.3 cm (in. diameter) and 19.5 cm (in. long).

This packaging has enough space to have the whole electronic circuitry which is integrated or the whole IF amplifier circuit board which is made from discrete components.

C-4 The Design of the First IF Amplifier and Voltage Regulator

The intermediate frequency was determined as 120 MHz in Chapter III. The IF bandwidth is 40 MHz. The circuit diagram of the outdoor unit and the parts list are summarized in Figs. C-5 and C-6 and in Tables C-2 and C-3.

C-5 The Design of the Indoor Unit

Figures C-7 through C-9 show the circuit diagrams of the second IF amplifier board and the limiter-discriminator board and the

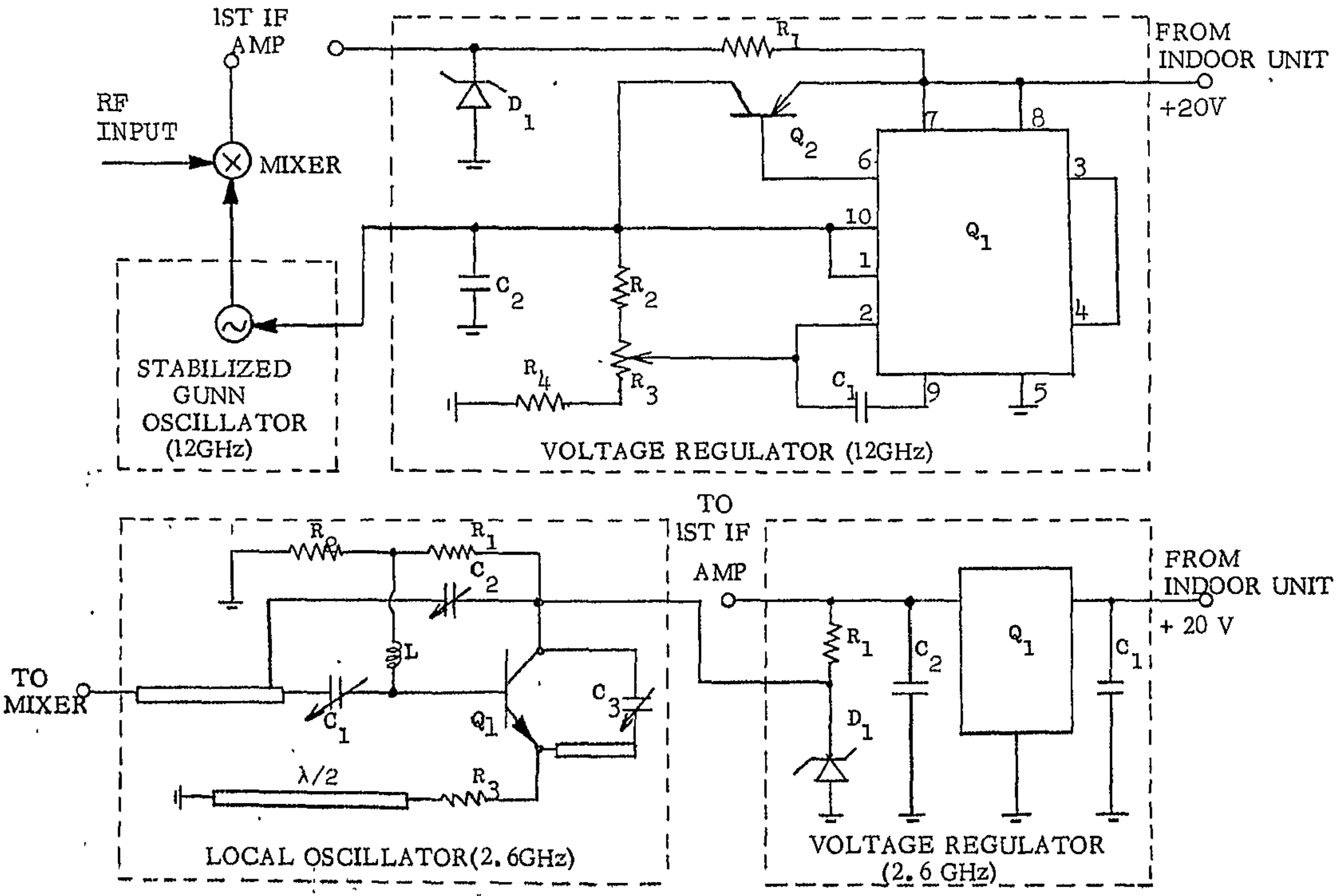


Figure C-5 Circuit Diagram of the Front-end and the Voltage Regulator of the Outdoor Unit

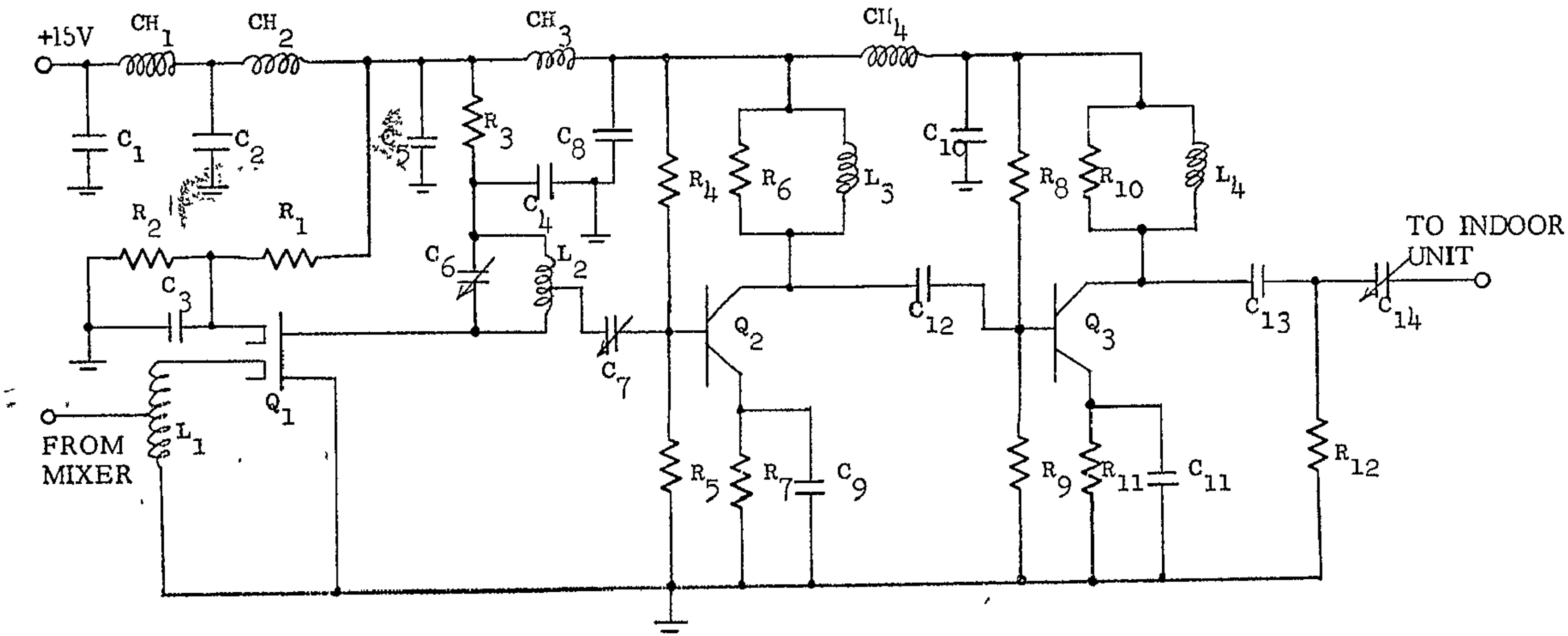


Figure C-6 First IF Section of the Outdoor Unit

Table C-2

2.6 GHz Outdoor Unit Parts List

Feed

a. Feed Cylinder
b. 6 Stubs and Mounting Nuts

Housing Assembly

a. Mixer Housing Body
b. Mixer Components
   1. Crystal Holder, Cold
   2. Shorting Rod
   3. Crystal Holder, Hot
   4. Coaxial Assembly
   5. Coaxial Insulator
   6. Collar Nuts
   7. Plugs
   8. Crystal
c. Local Oscillator Filter
   1. Filter Cover
   2. Filter Insert
   3. Local Oscillator Probe
   4. Filter Probe
   5. Teflon Bushing
d. Local Oscillator Housing
e. 1st IF Housing
f. Cover for LO and IF
g. Plugs and Jacks

Local Oscillator

R_1: 6.8 K Ohms

R_2: 2.7 K

R_3: 100

Q_1: MT 1061A, Fairchild

L : 4T, #18 Wire, 0.25" DIA.

C_{1-3}: 1 - 10 pf, Trimmer

Circuit Board

Regulator

Q_1: #7815, Fairchild

$C_{1,2}$: 0.33 μf

R_1: 100 Ohms

D_1: 1 N 5237, Zener

Table C-2 (Cont.)

1st IF Amplifier

R_1	:	100 K Ohms
R_2	:	47 K
R_3	:	220
$R_{4,8}$	:	10 K
$R_{5,9}$	:	3.9 K
R_6	:	680
$R_{7,11}$	:	1 K
R_{10}	:	2.2 K
R_{12}	:	150
$C_{1-4,8-11}$	:	1000 pf
C_5	:	1 μf
$C_{6,7,14}$	:	9-35 pf, Trimmer
C_{12}	:	15 pf
C_{13}	:	36 pf
Q_1	:	3N1 59, FET, RCA
$Q_{2,3}$	:	2N 3563
L_1	:	7 Turns, 1/2" DIA, 7/8" Long #1 AIRDVX #508 or B&W 3002
L_2	:	4 Turns, 3/8" DIA, 1/2" Long #16 Wire tapped 1/2 turn from top
L_3	:	100 nh 14 T #22 ON 0.1" DIA
L_4	:	180 nh 19 T #24 wire ON 0.1" DIA

Table C-3

12 GHz Outdoor Unit Parts List

Feed

a. Multi-Mode Horn
b. Polarizer
c. High-Pass Filter
d. Circular Waveguide - Coaxial Transition

Mixer

a. Diodes Hewlett-Packard
b. Circuit Rat-Race Hybrid
c. Housing Brass

Gunn Oscillator

a. Gunn Flange, Fairchild GOX-107
b. Main Cavity, Aluminum
c. Stabilizing Cavity, Invar Cylinder, Invar Plate and Brass Plate
d. Rectangular Waveguide - Coaxial Transition

Cover

Aluminum Cover Screws

1st IF Amplifier

Same as the 2.6 GHz Outdoor Unit

Voltage Regulator

R_1: 220 Ohms
R_2: 680
R_3: 1 K
R_4: 2.7 K
C_1: 470 pF
C_2: 15 nF
Q_1: A723C, Fairchild
Q_2: D42C
D_1: 1N5245

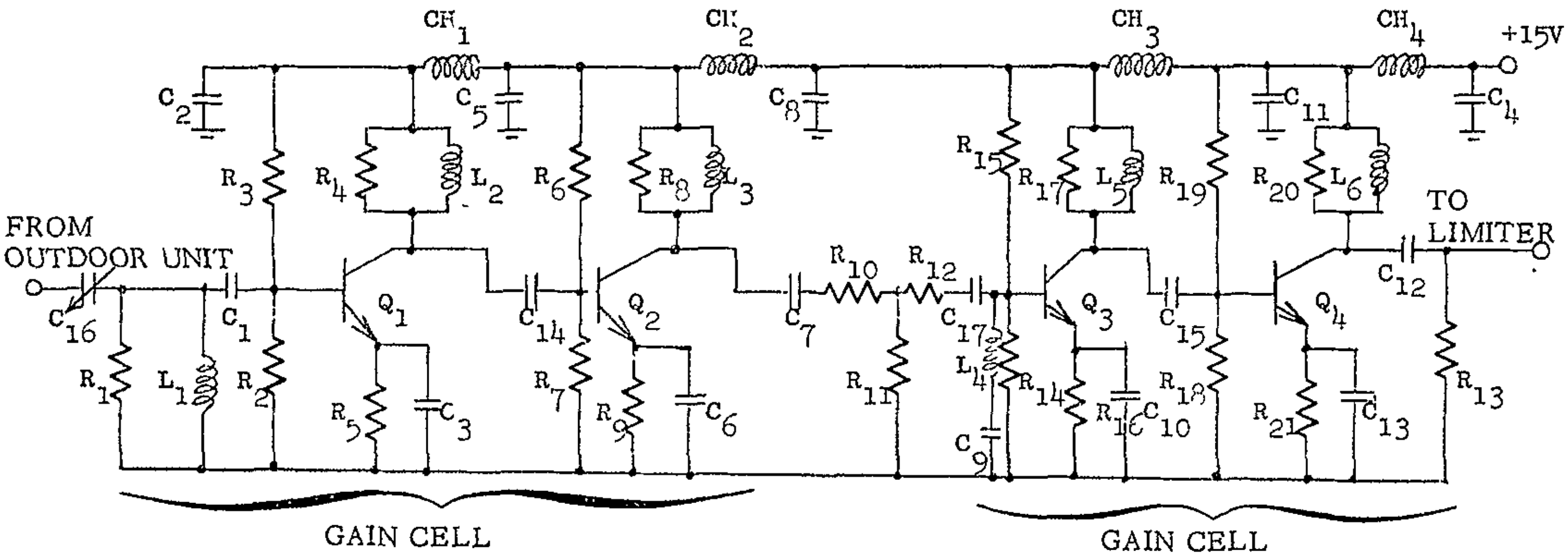


Figure C-7 Circuit Diagram of the Second IF Amplifier Circuit of the Indoor Unit

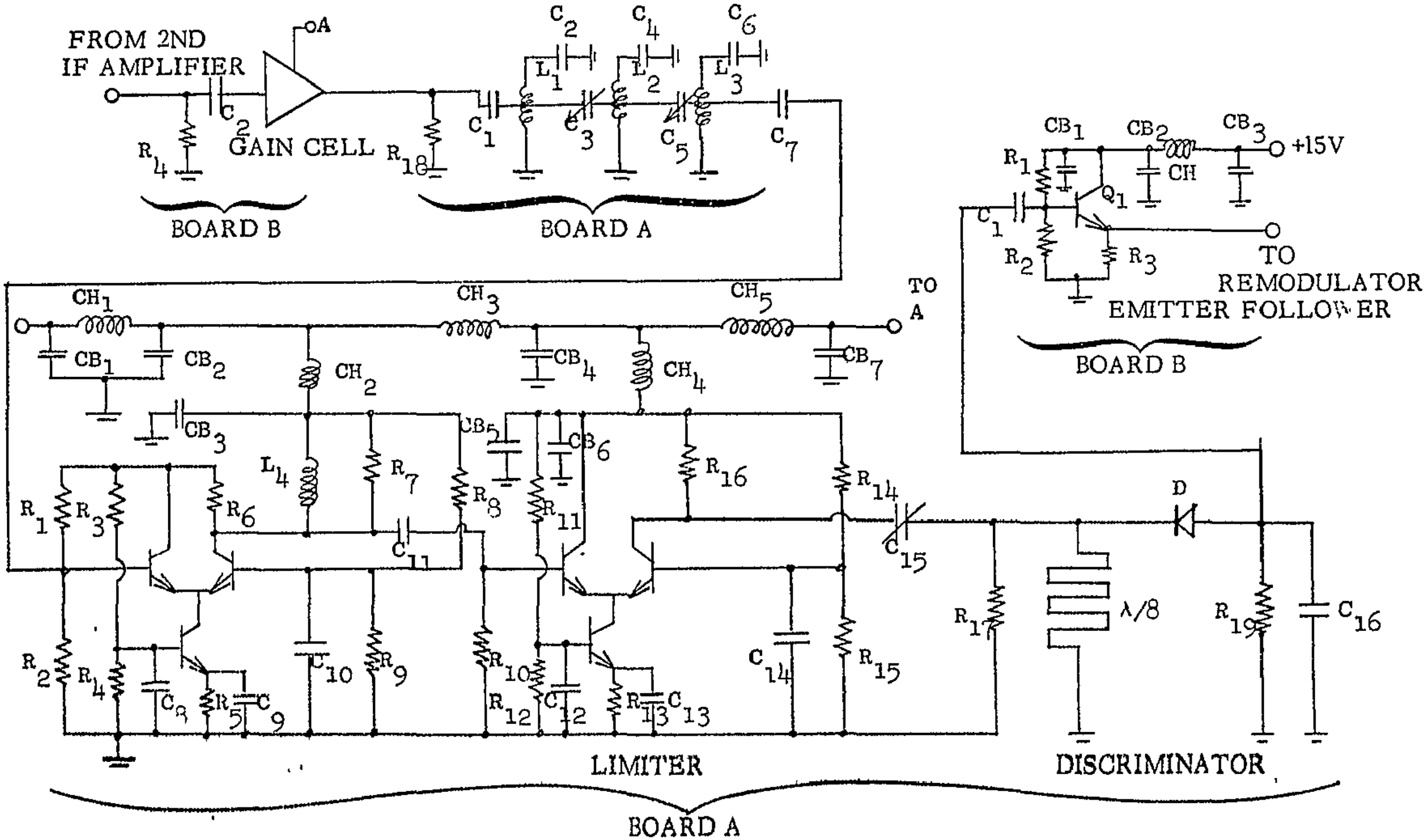


Figure C-8 Circuit Diagram of the Limiter and Discriminator of the Indoor Unit

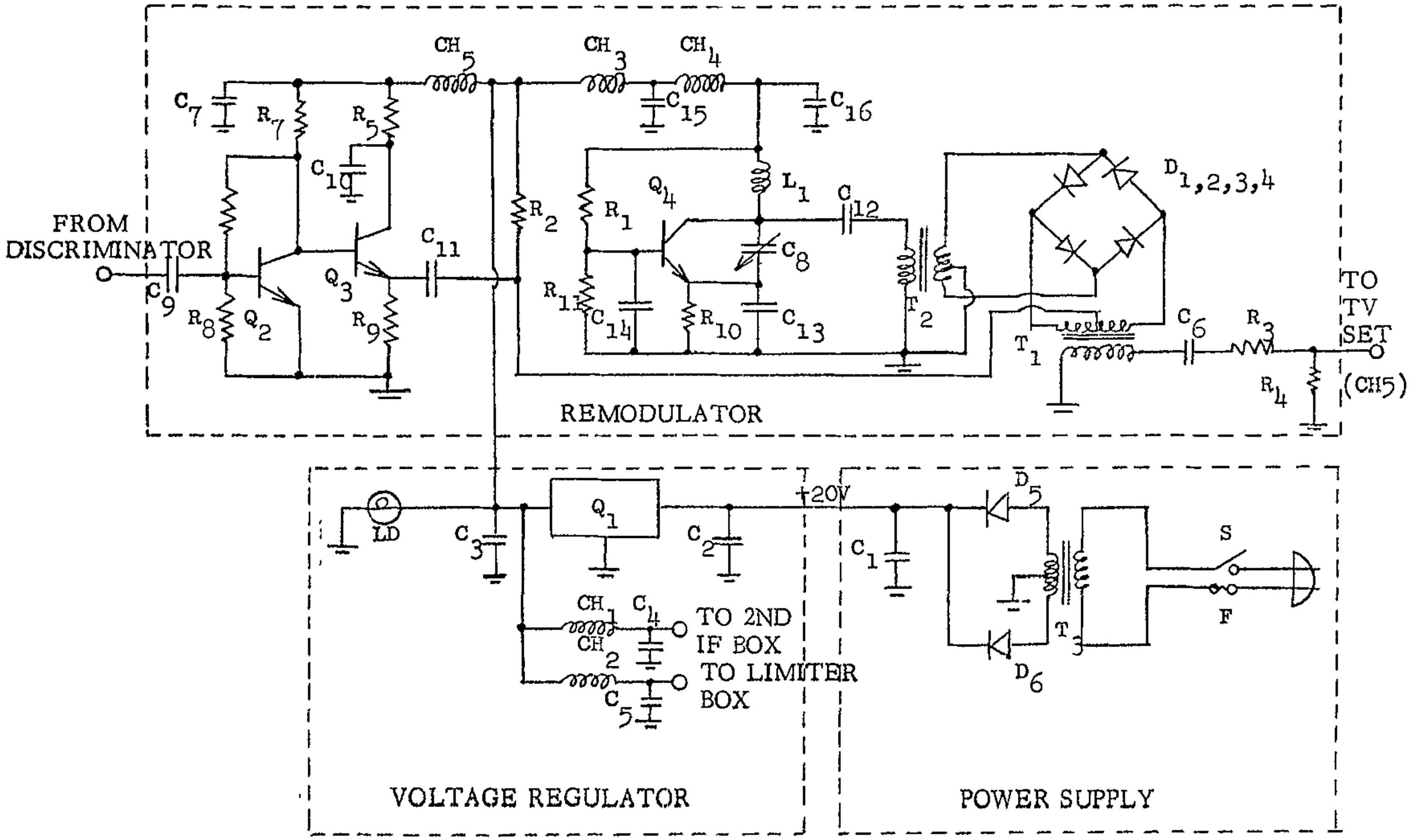


Figure C-9 Circuit Diagram of the Remodulator, Voltage Regulator and Power Supply of the Indoor Unit

remodulator-voltage regulator board, respectively. The parts list is shown in Table C-4. The remodulator circuit is only for the home reception.

Table C-4

Indoor Unit Parts List

2 ND IF Amplifier

$R_{1,13}$	:	100 Ohms
$R_{2,7,14,18}$	:	3.9 K
$R_{3,6,15,19}$	:	10 K
$R_{4,17}$	:	680
$R_{5,9,16,27}$	:	1 K
$R_{8,20}$	:	2.2 K
$R_{10,12}$	:	10
R_{11}	:	150
C_{1-13}	:	1000 pf
$C_{14,15}$	:	15 pf
C_{16}	:	9-35 pf Trimmer
C_{17}	:	36 pf
$L_{1,4}$	:	60 nH
$L_{2,5}$	:	100 nH, 5 T, #32 wire on 100 K resistor
$L_{3,6}$	:	180 nH, 7 T, #32 wire on 10 K resistor
Q_{1-4}	:	2 N 3563
CH_{1-4}	:	1.5 µH

Limiter Module
(BOARD A)

$R_{1,7,8,14}$	:	10 K Ohms
$R_{2,9,10,15,17}$	:	12 K
$R_{3,11}$	:	3.3 K
$R_{4,12}$	:	3.9 K
$R_{5,13}$	:	1 K
R_6	:	270
R_{16}	:	330
$R_{18,19}$	:	56
C_1	:	100 pf
$C_{2,4,6}$	:	27 pf
$C_{3,5}$	:	24 Normal or 9-35 pf Trimmer
$C_{7,16}$	:	100 pf

Table C-4 (Cont.)

C_{8-14}	:	1000 pf
C_{15}	:	9-35 pf Trimmer
L_{1-3}	:	65 nH, 4 T, #22 AWG, 3/16" DIA
L_4	:	3 T, #20 AWG, 0.25" DIA
IC	:	RCA 3049
D	:	1N 277
CB_{1-7}	:	1000 pf
CH_{1-5}	:	1.5 nH

(BOARD B)

R_1	:	2.2 K Ohms
R_2	:	2.7 K
R_3	:	82
R_4	:	56
C_1	:	50 μf
C_2	:	1000 pf
CB_1	:	1000 pf
CB_2	:	50 μf
Q_1	:	2 N 3563
CH	:	1.5 nH

Remodulator/Voltage Regulator/Power Supply

R_1	:	10 K Ohms
R_2	:	8.2 K
R_3	:	1.2 K
R_4	:	560
R_5	:	150
R_6	:	12 K
$R_{7,8}$	:	1.8 K
R_9	:	220
R_{10}	:	1 K
R_{11}	:	2.7 K
C_1	:	5500 μf, 25 V

Table C-4 (Cont.)

$C_{2-7,14-16}$	:	100 pf
C_8	:	9-35 pf Trimmer
$C_{9,10}$	:	22 µf
$C_{11,13}$	:	100 µf
C_{12}	:	3 pf
$T_{1,2}$	:	Micrometals Core, T25-12, 10 T, #32
T_3	:	F - 91 X Triad
D_{1-4}	:	1 N 4005
$D_{5,6}$	:	1 N 4005
Q_1	:	GE D 42 C2
Q_{2-4}	:	2 N 3563
CH_{1-4}	:	1.5 µH
CH_5	:	270 µH
L_1	:	Micrometals Core, T 25-12, 7T, #32
S	:	Switch
F	:	Fuse, 0.5A
LED	:	HP 5082

Appendix D

Multi-Channel TV Transmission

D-1 Introduction

So far, there has been a discussion of ground-terminal optimization and cost sensitivity analysis for one video and one audio channel TV broadcasting from satellites to ground terminals. The technical and economical aspects for multi-channel transmission which will more than likely be used in the future will be discussed in Appendix D. The multi-channel transmission is possible either by a multi-carrier system (multiple FM) or by a one-carrier double modulation system (multiple AM-FM). Communication system analysis described in Chapter II is extended to analyze the signal bandwidth and SNR for the multi-channel transmission.

D-2 Multi-Carrier System (Multiple FM)

Each channel has one main carrier and occupies identical signal bandwidth BW_{rf}. Therefore, the total bandwidth BW_{total} is given by

$$BW_{total} = 1.1 \cdot n \cdot BW_{rf} \qquad \text{(D-1)}$$

where a 10% guardband is assumed. The signal block diagram is shown in Fig. D-1.

The modification of the receiver for this multiplex is rather simple. Under such uncontrolled variables for the decision making that:

(1) the tuning mechanism has to be installed in the indoor unit

(2) spurious radiation must be within the FCC limit

(3) total frequency stability must be within ± 3 MHz.

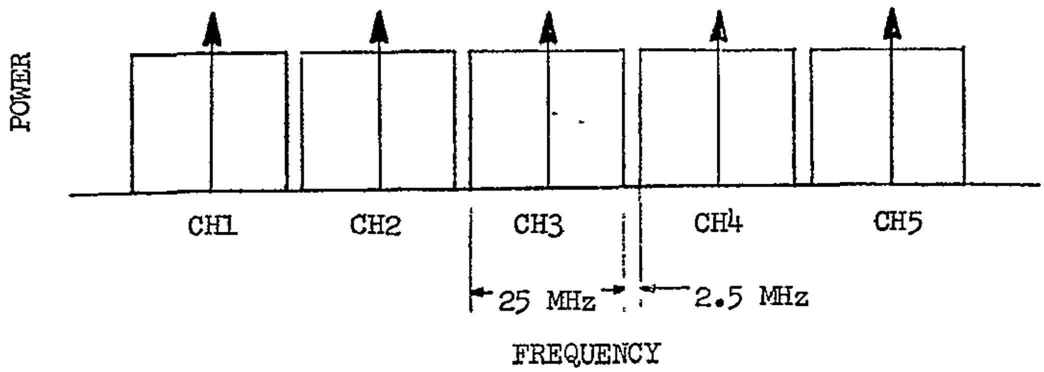


Figure D-1 Signal Arrangement of the Multiple FM System (in case of 5 channels)

A double conversion system as shown in Fig. D-2 is proposed as the best alternative. The circuits to be added are:

(1) a channel selector in the indoor unit required from the uncontrolled variable 1

(2) a UHF varactor tuner in the outdoor unit required from the uncontrolled variable 2

(3) an AFC circuit in the outdoor unit required from the uncontrolled variable 3

All these additional circuits are made by low cost parts which are commercially available in the TV industries.

The receiver terminal cost vs number of channels is shown in Fig. D-3. The satellite transmitter cost is approximately proportional to the number of channels n.

D-3 Single-Carrier System (Multiple AM-FM)

In Chapter II, a discussion of AM-FM double modulation system was given and it concluded that the AM-FM system has poorer FM improvement and is less promising for satellite broadcasting. But taking into

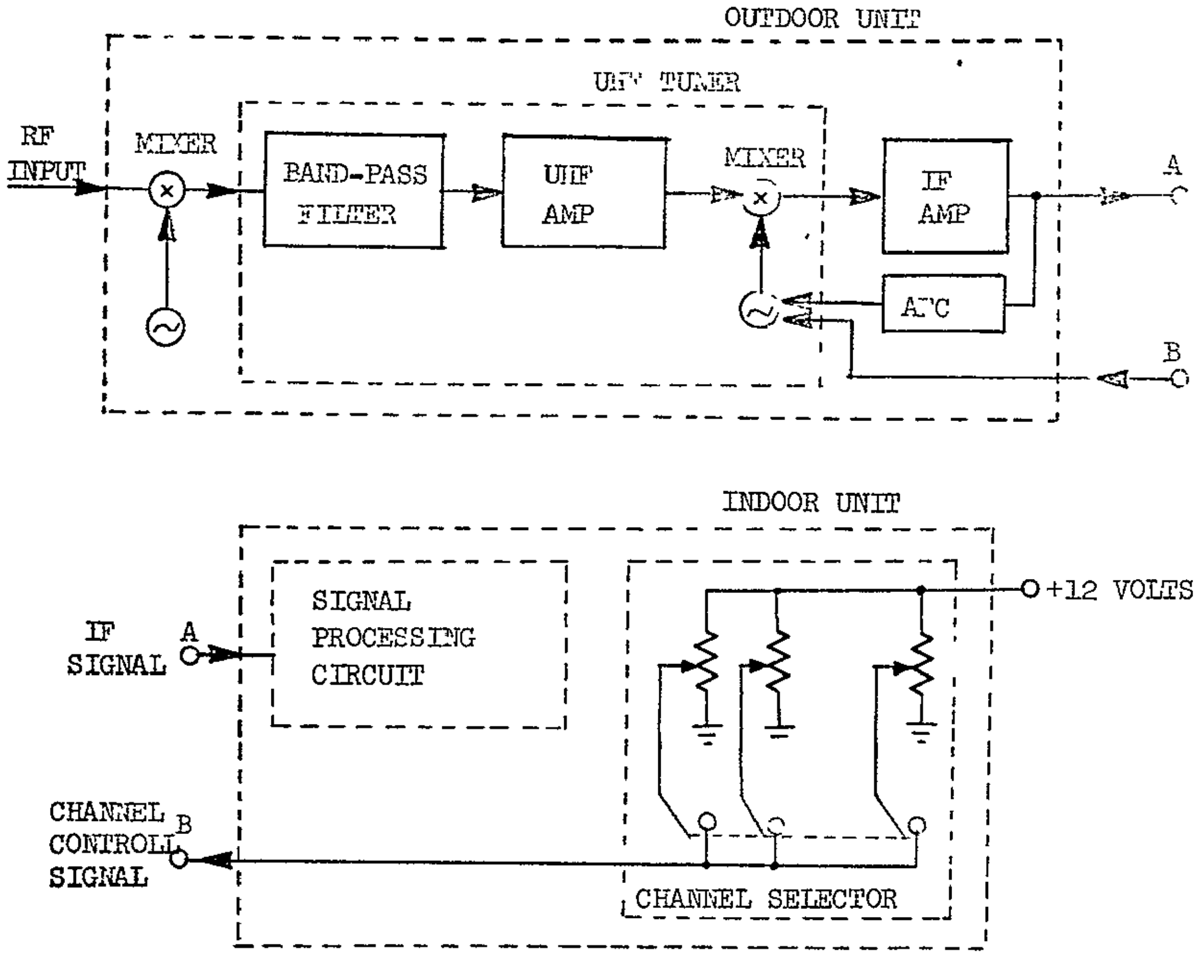


Figure D-2 Modified Circuit Block Diagram for the Multiple FM System

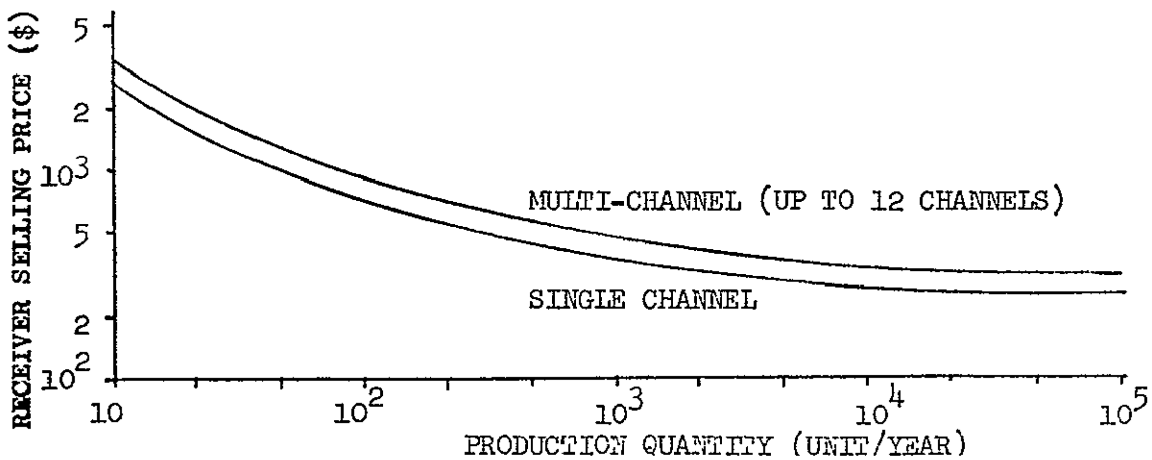


Figure D-3 Receiver Unit Cost vs Annual Production Quantity

consideration that the receiver circuit needs no modification except broadening the bandwidth of the indoor unit circuits, it is worth analyzing for the case of multi-channel transmission, also.

The modulating signal for multiple AM-FM transmission is a chain of VSB-AM_{video} -FM_{audio} signals arranged every 6 MHz as shown in Fig. D-4. The required bandwidth is, as stated in Chapter II

$$BW_{rf} = 2(f_{vn} + \Delta f) \tag{D-2}$$

where f_{vn} is the video subcarrier frequency of the highest channel. The FM improvement factor from Chapter II is also given

$$FMI = \frac{3}{2} \frac{(\Delta f)^2}{f_{hn}^{\ 3} - f_{ln}^{\ 3}} \left(\frac{m}{1 + m}\right)^2 BW_{if}. \tag{D-3}$$

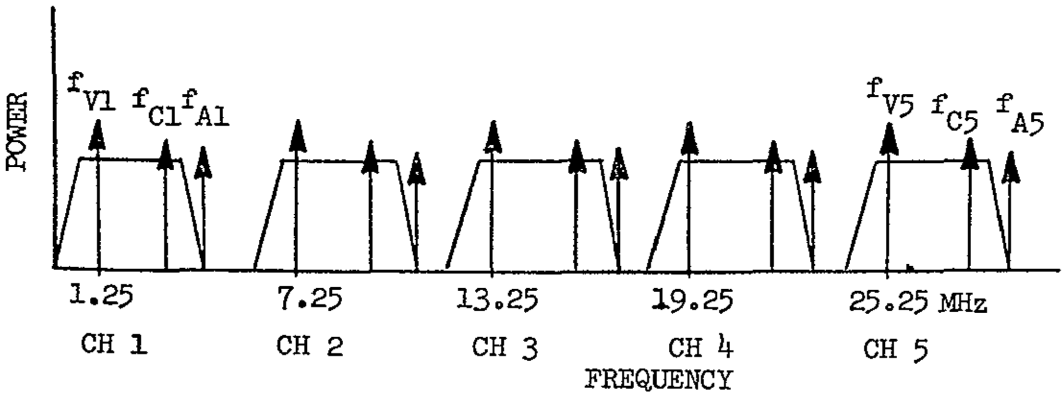
The fundamental relationship between communication parameters is found in Chapter II,

$$SNR = CNR \cdot FMI \cdot PPW \cdot NWF \tag{D-4}$$

where NWF is 5.5 dB because the noise is almost flat for larger n. Then, in order to get 45 dB or 42 dB of SNR and 11 dB of CNR,

$$FMI = 19.5 \text{ or } 16.5 \text{ dB} \tag{D-5}$$

respectively, is required. Figure D-5 shows the necessary bandwidth per channel for the multiple AM-FM system with comparison to the multiple FM system.



DEFINITIONS:

f_{Vn}; VIDEO SUB-CARRIER FREQUENCY FOR CHANNEL n
f_{Cn}; COLOR SUB-CARRIER FREQUENCY FOR CHANNEL n
f_{An}; AUDIO SUB-CARRIER FREQUENCY FOR CHANNEL n

Figure D-4 Modulation Signal Configuration for Multiple AM-FM

AM-FM requires approximately 2 to 2.5 times as much bandwidth as FM does, in order to get the same FM improvement or 8 dB more EIRP to broadcast the same quality by the same bandwidth as FM. This fact is quite discouraging although it has a big merit factor in that the same receiver as was proposed can be used for the AM-FM multiples without any modification but broadening the circuit bandwidth of the indoor unit.

Taking into consideration that the modification of the receiver of the multiple FM system is also fairly low in cost and that, on the other hand, the multiple AM-FM system may have serious difficulty in intermodulation and cross-modulation problems, it is concluded that the multiple FM system is superior to the multiple AM-FM system for the multi-channel satellite TV broadcasting.

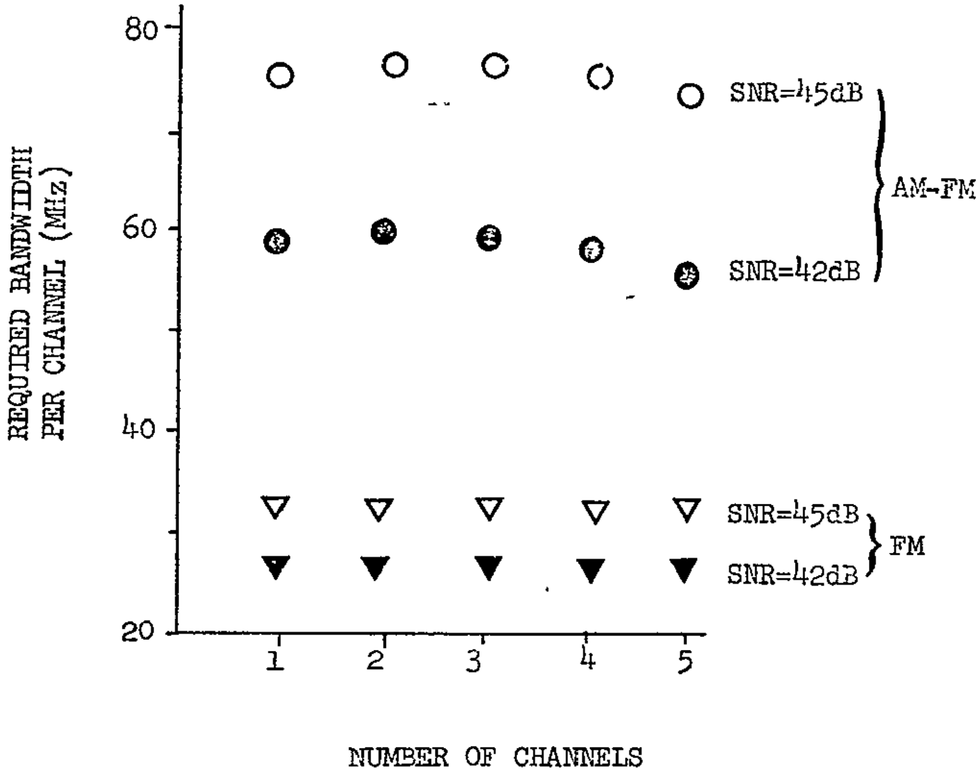


Figure D-5 Required Bandwidth per Channel vs the Number of Channels

Appendix E

2.6 GHz Receiver Modifications

E-1 Outdoor Unit

E-1-1 Polarizer

The polarizer was changed from a two lumped element item to a six element stub system to provide a good VSWR and good axial ratio over the entire band from 2.5 to 2.7 GHz. The older two-element system provided good axial ratio only at the desired received frequency of 2.62 GHz. This new polarizer is designed to be removable and also can be rotated about the axis of the feed cylinder so that either left or right hand circular polarization can be obtained.

E-1-2 Local Oscillator

The local oscillator was changed from a crystal controlled source to a direct transistor oscillator. The new oscillator uses a MT 1061A transistor made by Fairchild which cost less than $5 in moderate quantities. The device has been made to oscillate in a fundamental mode by integrating the package parasitics into the circuit design. Operation with this device normally was limited to 1800 MHz previously. The principal savings have been in the amount of time required to set the oscillator at the proper frequency and proper power output, and with the improved stability of operation over a larger temperature range. This oscillator exhibits a plus and minus 2 MHz variation in frequency over temperature range for -5^{o}F to $+140^{o}$F. For operation at the lower temperatures, it is suggested that a heater be incorporated in the oscillator design. A voltage regulator was added to the outdoor unit to help stabilize the oscillator frequency over temperature

and to eliminate interference from entering the system on the power cables.

E-1-3 Packaging

The local oscillator and the regulator are packaged in a small 2 x 4 inch bud die cast chassis box. The IF amplifier is housed in a similar box and both boxes are covered with a simple sheet metal weather tight covering. This simply replaces the die cast box of the original prototype. No other changes were made to the mixer or filter insert.

E-2 Indoor Unit

The circuitry for the indoor unit is housed in a simple two piece sheet metal box 4 x 6 x 8 inches. Die cast bud chassis boxes are used to house the individual circuit boards which is the second IF amplifier, the limiter discriminator, and the remodulator. The power supply components are attached directly to this larger chasis box. Modifications to the signal processing circuitry for the indoor unit follow.

E-2-1 IF Amplifier

The IF amplifier has been redesigned using a computer aided optimization program called OPTINET. The result is that no transformer or inner stage adjustments are necessary. Adjustable matching networks are provided at the system inputs and outputs, however, to eliminate reflections on the various cables, especially those which

could occur on the cable length between the indoor unit and the outdoor unit. Threshold test indicate that there was insufficient gain in the IF amplifier chain and so an extra gain cell was added inside the demodulator box.

E-2-2 IF Filter

The realization for the IF filter has been changed to a three pole synchronously tuned type of filter with adjustable coupling capacitors only. This has proven sufficient to provide the equalization and bandpass adjustments necessary for demodulation of the black and white video signal.

E-2-3 Limiter-Discriminator

The limiter discriminator was intended for use as a black and white signal demodulator and tests indicate that this is the extent of its capability. The only changes made to it were minor circuit modifications. The transmission line discriminator was shortened and was incorporated directly on the limiter board. An emitter follower circuit was added to the output at the detector to provide the low impedance necessary for use with a de-emphasis filter. The de-emphasis filter (according to the CCIR standard) was added using a separate box to facilitate testing.

E-2-4 Remodulator

Only minor changes in component values were necessary for the video amplifier and the remodulator circuitry. A 20 dB pad was

added at the output to set the impedance to 300 ohms balanced, and to prevent overdriving the television sets.

E-2-5 Power Supply and Regulator

Fixed output IC voltage regulators have recently become available and have been incorporated into the indoor unit. This is because these devices are now less expensive than the sum of the discrete components used to make them. A larger filter capacitor is needed for the indoor unit intended to operate with the 12 GHz outdoor unit because of the large current requirement of the Gunn diode oscillator.

E-3 Antenna

A 5 ft. to 10 ft. parabolic antenna is installed outside with the described outdoor unit at its focal point. Side-lobe suppression of the antenna is as important as high efficiency in order to avoid the interference from other satellites. Low sidelobe techniques have been investigated and are discussed in [33].

E-4 Improvements for Higher G_r/T_s

Recent advances in low-noise transistor device technology will make it possible to obtain system noise figures of 4.5 dB or better for only a modest increase in overall system cost. Such an improvement in noise figure will result in an overall increase in system sensitivity of more than 4 dB. This means that much smaller transmitters will be needed in the satellites to provide service to direct receivers. The

cost for obtaining reasonable gain at microwave frequencies such as 2.5 GHz is becoming so low that it is feasible to consider a receiver design which discriminates the signal at 2.5 GHz without resorting to an intermediate frequency or a local oscillator. Hewlett-Packard has proposed such a system for the ATS-F Health/Education experiment being sponsored by the Department of Health, Education, and Welfare. They have quoted a cost of $850 per receiver in quantities of 500 units. This is only slightly more expensive than our prediction. Use of such preamplifiers would necessitate a slight change in the feed-polarizer design. One could continue to use the waveguide feed polarizer and mount the preamplifier directly at the feed in a housing designed especially to hold it, but because it is known that the noise figure of such preamplifiers degrades severely as the temperature is increased above the ambient room temperature it may be advisable to mount that preamplifier behind the antenna. If this is the case, then a feed such as a helix would be more suitable since the cable that drives the helix could be run down the center supporting rod to the apex of the dish and the electronics package could be mounted just behind the dish out of the elements and away from the heat of the sun.

<u>Color Reception Requirements</u> - In order to receive color video transmissions it would be necessary to improve the limiter discriminator design. The most important parameter which must be imporved is that of amplitude linearity. In order to obtain the necessary improvements it would be necessary to increase the frequency response of the two limiter stages to approximately 180 MHz, in fact, the limiter response should be flat from the range of 50 to approximately 180 MHz. The

present configuration is not capable of achieving that kind of response and therefore, a new circuit will be needed. The phase linearity of the IF amplifier, the IF filter and the limiter has been adequate for color reception. However, if more than 1 audio subcarrier were needed to be transmitted, the phase linearity would not be adequate and equalization would be required, depending on the number of additional channels.

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THE FOLLOWING PAGES ARE DUPLICATES OF ILLUSTRATIONS APPEARING ELSEWHERE IN THIS REPORT. THEY HAVE BEEN REPRODUCED HERE BY A DIFFERENT METHOD TO PROVIDE BETTER DETAIL